ULY 1959
OL. 2 • NO. 7

- Tannenwald — Spin Waves
- Swanson — Modulator Techniques for Gridded Klystrons and Traveling Wave Tubes
- Southworth — Graphical Analysis of Microwave Components, Part II

he microwave journal

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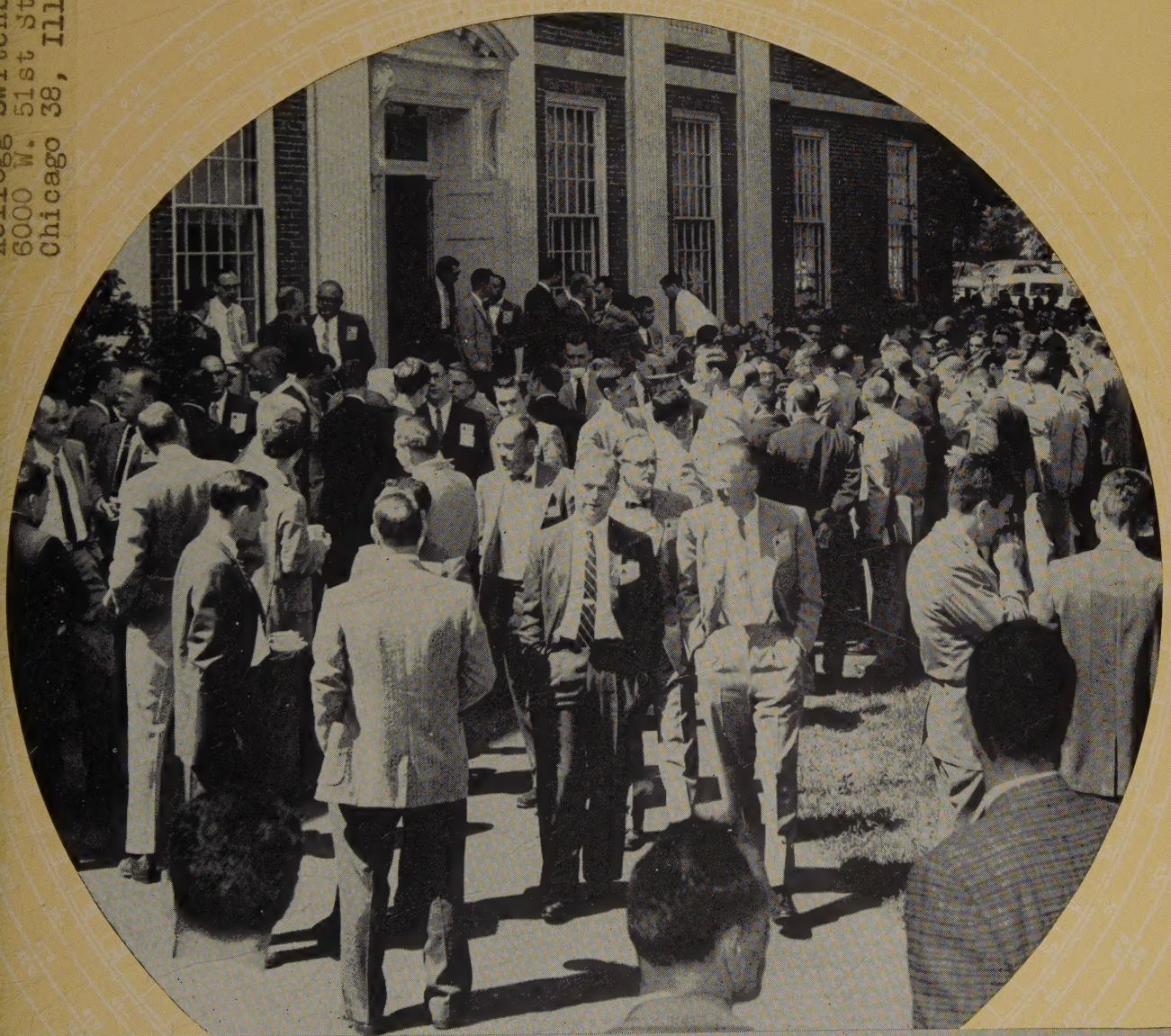

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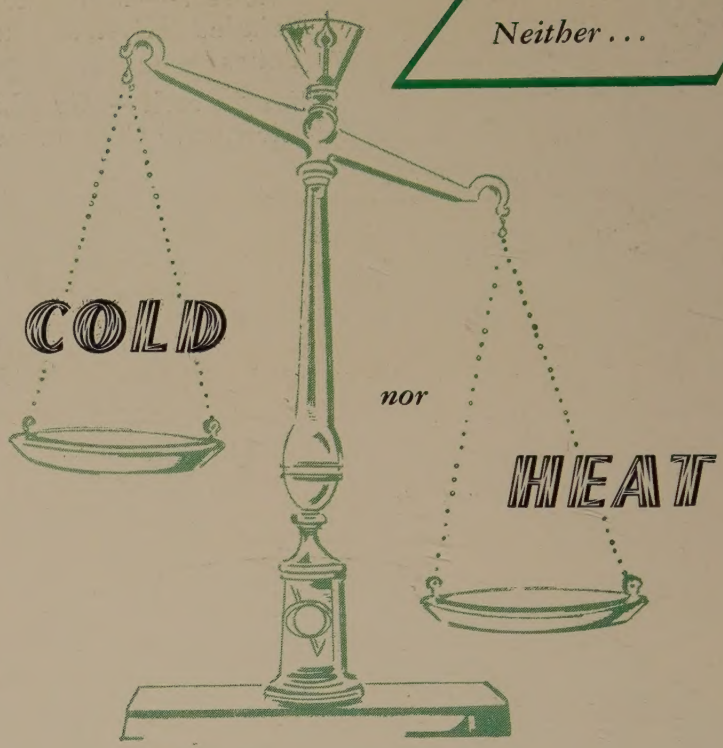

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these instruments
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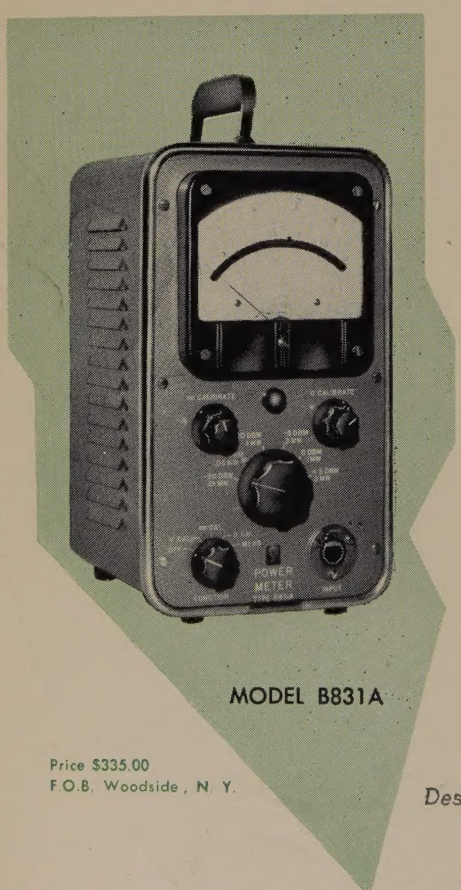

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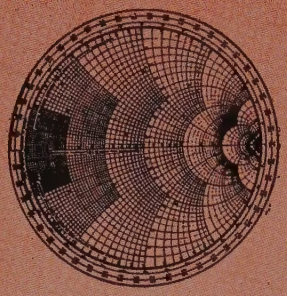

the microwave journal

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COVER PICTURE

Coffee Break at the 1959 PGMTT Symposium.
See page 52. — (Photo by Tore Anderson)

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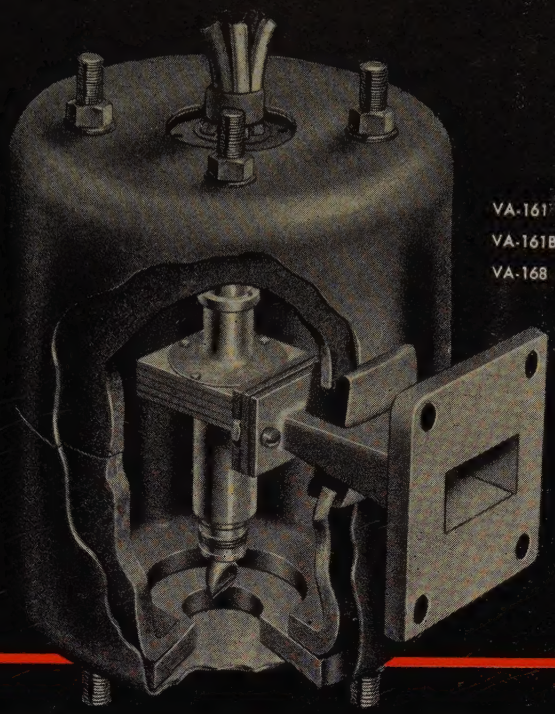

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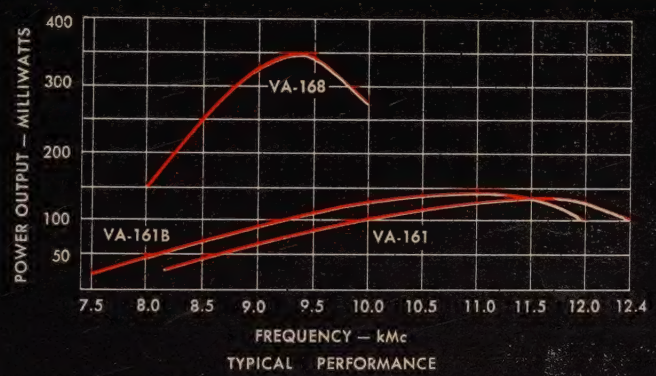

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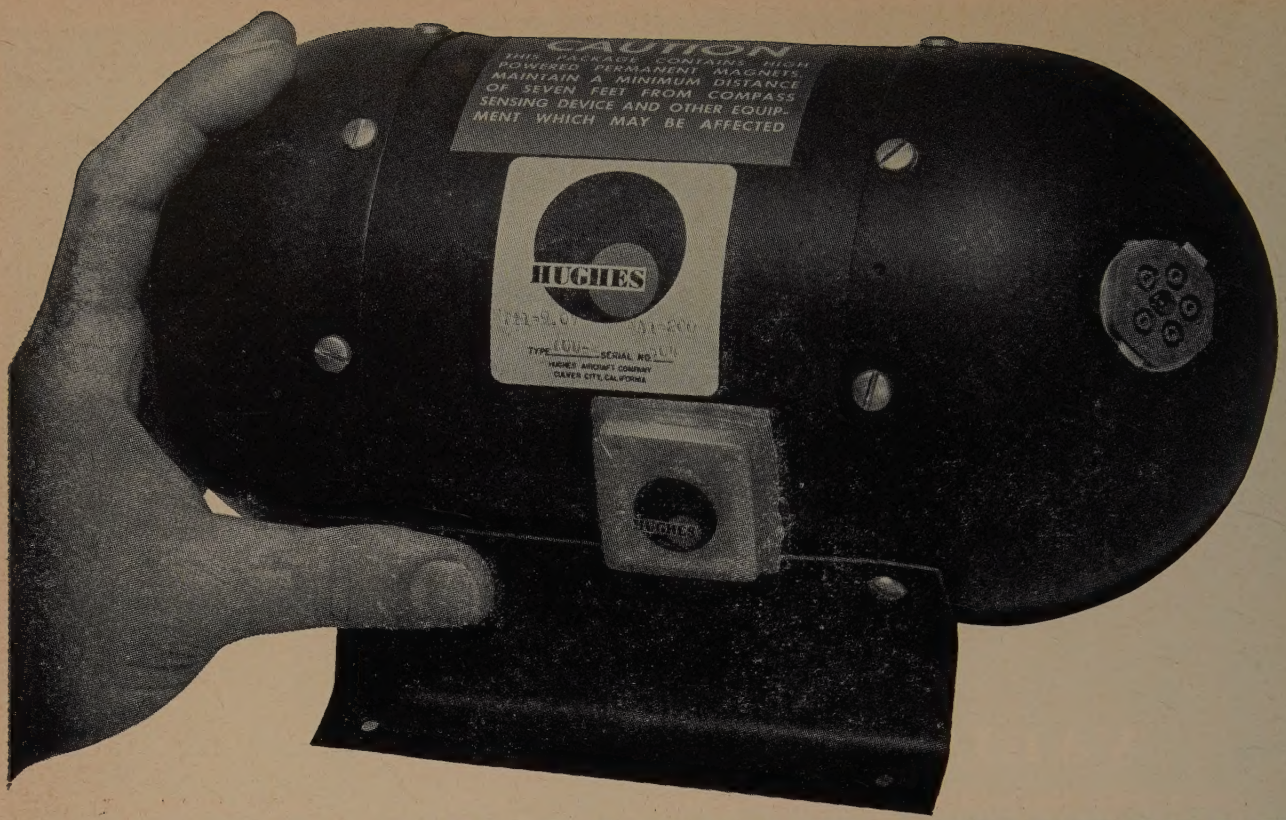
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For additional information please write: Hughes Products, Electron Tube Sales, International Airport Station, Los Angeles 45, California. For export write: Hughes International, Culver City, California.

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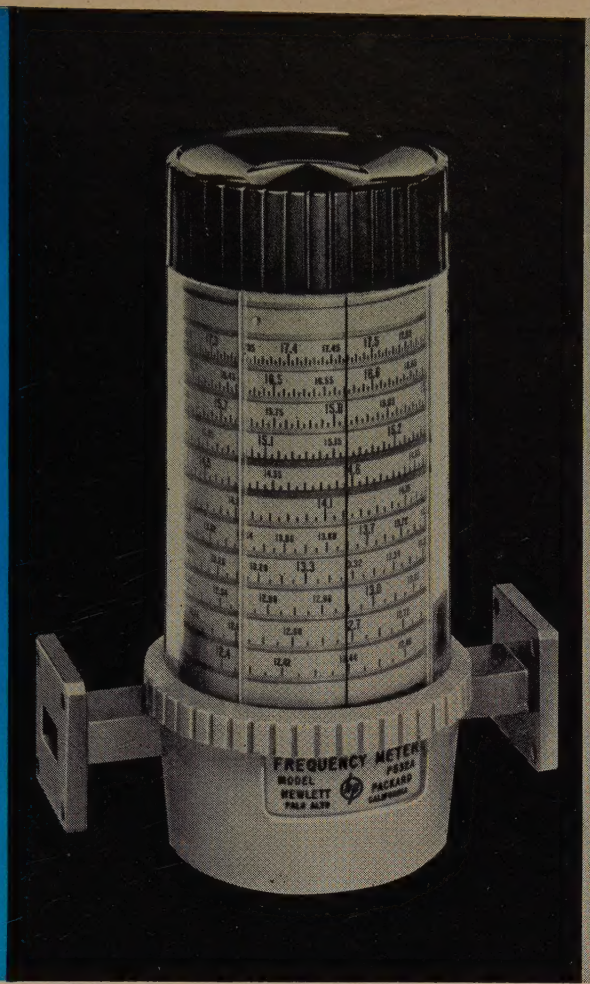

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For J and H band work (5.20 to 10.0 KMC) Model 530 Frequency Meters are offered (see Table). These instruments are similar to Model 532A except tuning is by a micrometer, and readings are converted to frequency by a chart on the instrument.

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J530A	0.1%	5.85 - 8.20	1½ x ¾	4	$120.00
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FROM THE EDITOR

One year ago the first issue of "the microwave journal" was published. Although the basic idea of the journal at that time was a sound one in which we all believed, the first issue was mailed out with a degree of, what shall we call it — hesitation? apprehension? or was it downright fear? We knew that we had in "the microwave journal" a medium that would fill a long-existing need in the technical publications field — but how about the hazards that face a new magazine in this highly competitive age? The same research that had revealed to us the need for a microwave journal also pointed out that x out of y magazines started had failed! Experience, we knew, would be almost a prime requisite of survival.

Now we have a year of survival, with pages of experience, some interesting, some boring, some amusing, but all valuable.

The year has been an exciting one. We have seen our original circulation of 12,000 grow to 20,000, and this despite careful pruning. We have received mail from all over the world expressing an interest in our journal.

Judging from our own limited observations, we have been well accepted by the technical people in the microwave industry. This technical acceptance is being followed by an increasing acceptance by advertising people, who, after all, are making it possible for us to continue to send you the magazine.

Our authors have been some of the most capable people in the field. We are grateful for the contributions of Dr. Southworth, Dr. Ginzton, Dr. Heffner, Dr. Ewen, Dr. Bowness and many others.

One of the important factors in the success of the journal to date has been the interest and work of our Associate Editors and Associates. In addition to their written contributions, they have been extremely active in areas of editing, procurement of papers, and the determination of editorial policies. Their role in this venture has gone far beyond that of being honorary names on the masthead. Without them and their efforts, the journal would still be fighting to publish its first issue.

In the past year we have managed to maintain the strict editorial policies set down long before the first issue was published. These policies can be expressed most simply by the phrase "quality of technical content without compromise." At best this is a difficult task. We have tried to maintain this policy in spite of tugs of expediency and other pressures. Any deviation from this rule has been primarily due to our own lack of experience.

With more sophistication and armed by the lessons learned in our freshman year in the publishing business, we will try to avoid future pitfalls and in so doing, we hope to bring you a more polished magazine. We will, however, never lose sight of the fact that it is the Microwave Industry that makes "the microwave journal" possible, and not vice-versa. We hope, as the publisher puts it, to serve the microwave industry as a forum and a marketplace.

One final note about the publisher. It is certainly a pleasure to work with a man who responds so readily to threats of physical violence.

Theodore S. Saad

IN THE NEXT ISSUE

The August issue of "the micro wave journal will include the follow ing papers: "Printed Circuit Radi ators — The Sandwich Wire An tenna" by Walter Rotman and Nich olas V. Karas of AFCRC, "The De sign of Multi-Hole Coupling Ar rays" by Edward S. Hensperger o Waveline, and "Low Frequency Pro totype Backward-Wave Reactanc Amplifier" by D. I. Breitzer and H W. Sard of A.I.L.

The Business Editorial will be fur nished by William R. Hewlett, Ex ecutive Vice-President of the Hew lett-Packard Company.

REPRINTS

Reprints of the ENTIRE TECH NICAL SECTION of "the micro wave journal" may be obtained fron the publisher at the following rate — 1 to 5 reprints, twenty-five cent each, 6 or more, fifteen cents each Rates on reprints in large volum may be obtained on application t the publisher.

AFCRC SYMPOSIUM

A classified symposium on "Th Plasma Sheath, Its Effect on Com munication and Detection" will b sponsored December 7 and 8 in Bos ton, Mass., by the Electronics Re search Directorate of the Air Forc Cambridge Research Center.

Purpose of the symposium is t exchange both classified and unclass fied information on the problem o transmitting radio frequency energ through the plasma sheath surround ing a hypersonic vehicle, and th companion problem of reflection o radio frequency energy as related t radar detection.

Twenty-five minute papers are in vited. Information may be obtaine from Miss Alice Cahill, Air Forc Cambridge Research Center, Attn CRRD, Laurence G. Hanscom Fiel Bedford, Mass.

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A-15049*	1100 500	0.003 0.06	500 500	1000 1000	1250 1250
RCA-2039	1500	0.06	200	200	250
RCA-6952	2000	0.0018	425	600	1000
A-2344*	5000	0.003	1000	1000	1250
A-2349*	8000	0.003	200	200	250
A-2346*	10000 5000	0.01 0.06	450 450	450 450	600 600
A-15025*	27500	0.003	425	600	600

*RCA Developmental Type ‡For Prototype Design
■ At Peak of Pulse

Typical data on RCA Super-Power Tubes in hard-tube pulse-modulator service

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A-15034*	11,000	0.05

*RCA Developmental Type

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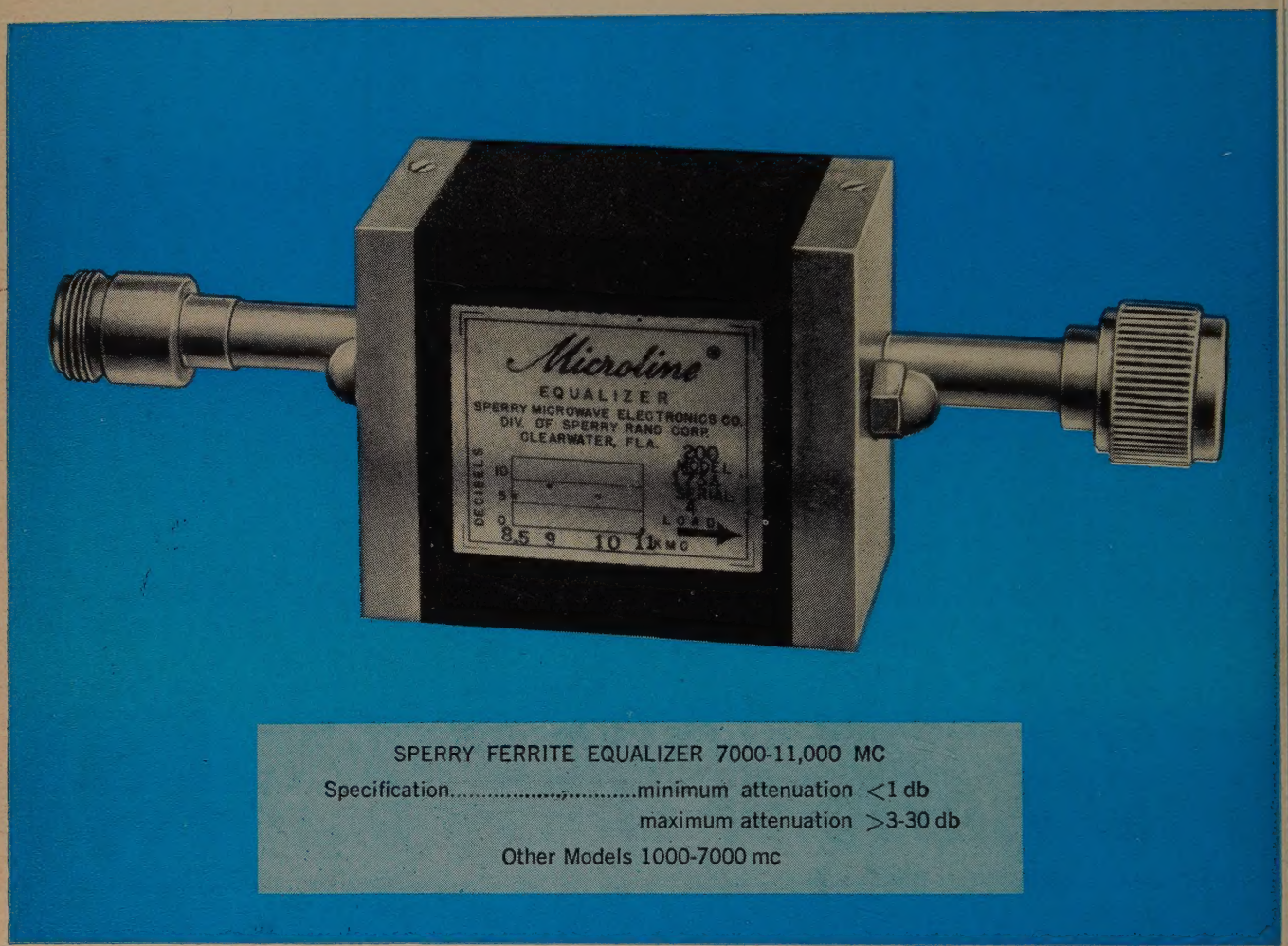

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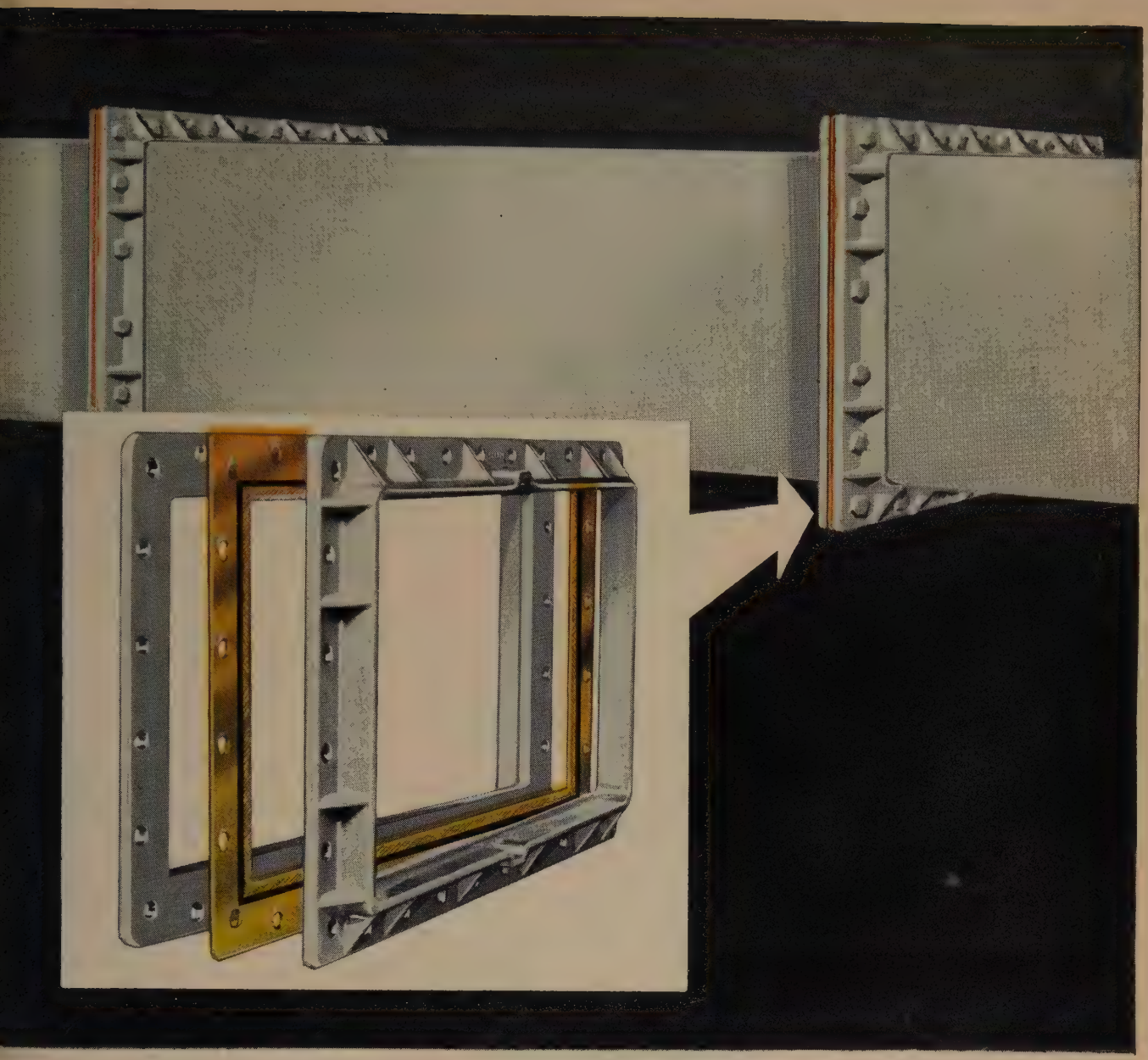

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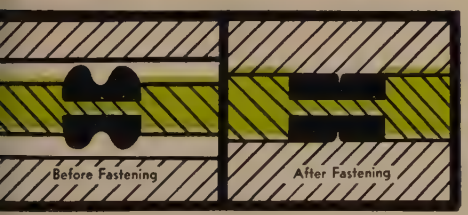

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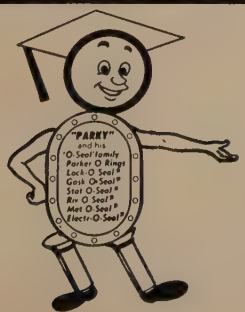

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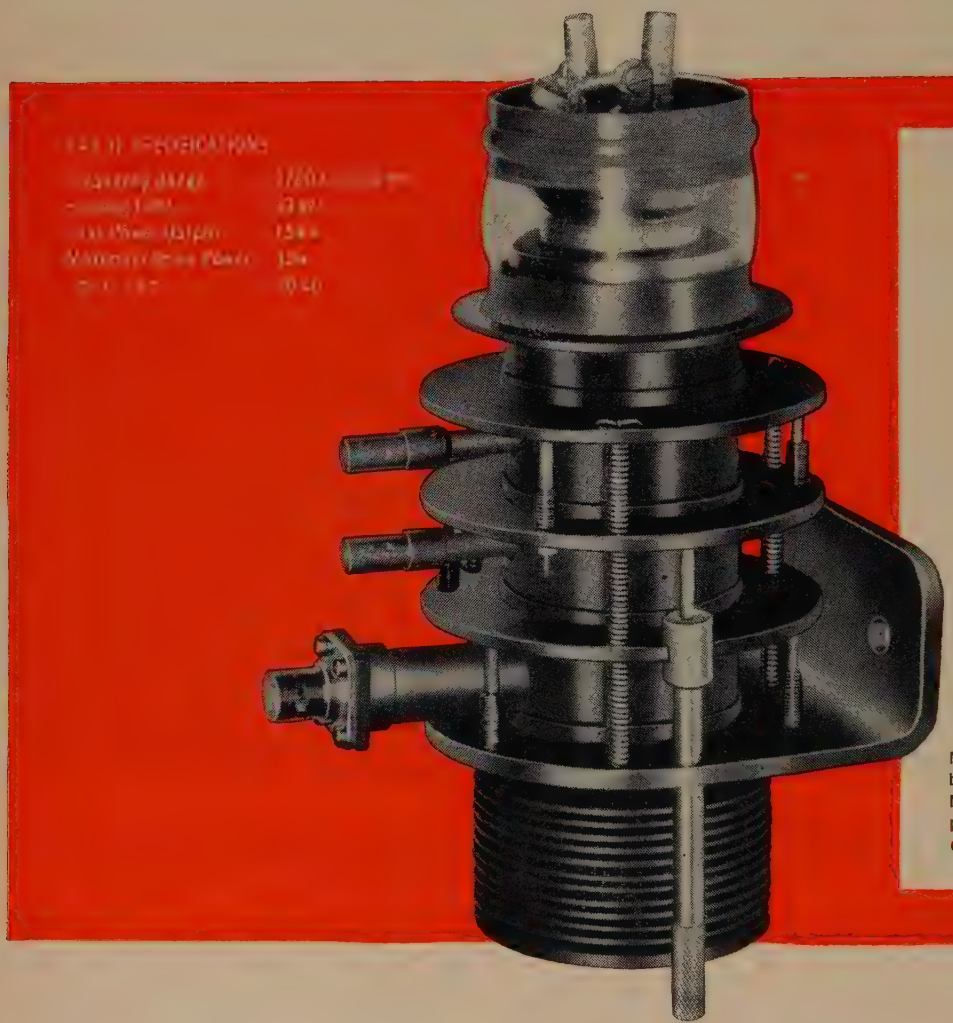

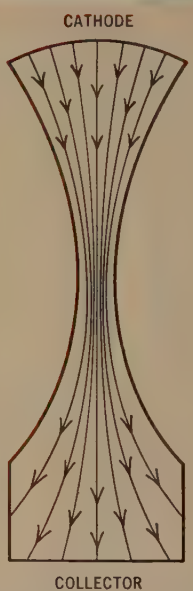

New Space Charge Focus principle of beam control is shown in diagram. New Sperry tube design utilizing this principle reduces size, weight, power consumption and cooling needs.

AVAILABLE FOR IMMEDIATE DELIVERY, Sperry's new S-band transmitting tube is a 3-cavity pulse amplifier of high gain and extra-long service life.

EXCLUSIVE SPERRY SPACE-CHARGE FOCUSING design eliminates heavy, cumbersome magnetic structures — a feature of prime importance in equipment design. Although the SAS-61 weighs only 6½ lbs., its sturdy construction withstands extreme vibration and environmental conditions.

MAIN APPLICATIONS for the SAS-61 are as an output tube in low-power radars, or as a driver for higher-powered klystrons in radar and linear accelerator systems. Its unusually long service life, however, makes it highly desirable for any application requiring 15 kw in the S-band. The SAS-61 with its internal tunable cavities is a *complete* microwave unit. No external equipment is required.

SPERRY CAN DELIVER SAS-61 tubes in quantity at once. Write for data sheet E-7A.

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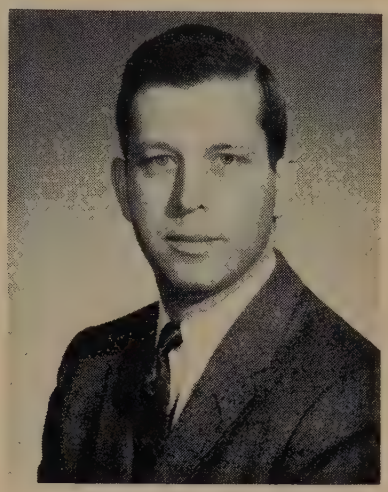
The introduction and systematization of frequency filters by G. A. Campbell and O. J. Zobel, among others, have resulted over a 40-year period in virtually countless filter applications of great importance, particularly in the audio- and lower radio-frequency ranges. Great industries have grown from the development of equipment containing filter circuits as essential elements, and in non-commercial forms frequency filters have made possible the communication and radar capabilities of the armed services. Although recognition of the importance of frequency filters has spurred an enormous amount of research, the results apply principally to the audio and lower radio frequencies where inductances, capacitances, and resistances may be treated as constant, lumped elements. At higher frequencies, where the dimensions of inductances, capacitances, and connecting leads become an appreciable fraction of a wavelength, these elements can no longer be assumed to be constant with frequency and concentrated at discrete points in space. At still higher frequencies, the lumped-element viewpoint fails altogether, and filter networks must be designed in terms of their transmission-line, waveguide, and resonant-cavity constituents. These two frequency ranges — one where lumped-element circuits must be compensated for their distributed nature and the other where distributed elements must be employed exclusively — have much in common and together represent the spectrum of interest to microwave engineers.

THE CURRENT STATUS OF MICROWAVE FILTER DESIGN*



BY SEYMOUR B. COHN

Stanford Research Institute
ASSOCIATE EDITOR — THE MICROWAVE JOURNAL

A FREQUENCY FILTER may be defined as a device that is capable of transmitting desired frequency signals and of rejecting undesired frequency signals between one or more generators and one or more loads. In the usual filter, the transmission property occurs with relatively low loss of signal power over one or more regions (known as pass bands) in the frequency spectrum, while the rejection property occurs with relatively high loss of signal power in one or more other regions known as stop bands. The various types of filters are usually named according to their insertion-loss-versus-frequency response characteristics. The more common types of response functions are illustrated in Figure 1, and the nomenclature will be clear from the typical curves. In each case a pass band is defined as a frequency region in which the insertion loss is less than some specified limit, and a stop band as a region in which the insertion loss exceeds a specified limit. A finite frequency range necessarily exists between adjacent pass and stop bands to accommodate the large change in insertion loss. A virtually infinite number of design possibilities exists for each type of filter depending upon the frequency and insertion-loss requirements placed on the pass, stop, and transition bands. The problem is made still more complicated for the designer in that he must choose from a number of available design methods. The most suitable method must be decided in each case from considerations of the type of response desired, the frequency range, the severity of performance requirements, and the economic conflict between development and manufacturing expenditures.

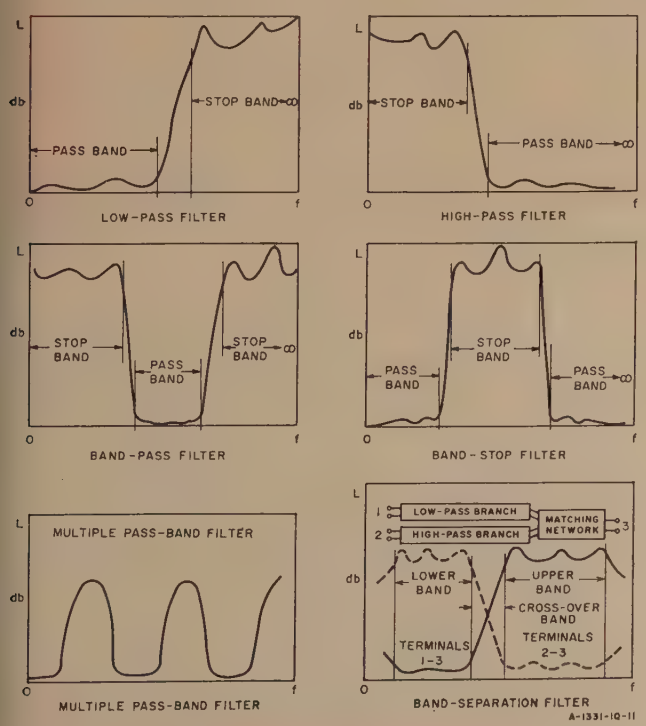


Figure 1—Insertion loss versus frequency response curve.

The Image-Parameter Method of Design

By far the most widely used method of filter design makes use of the so-called image parameters. Although this method evolved in the infancy of filter work, its simplicity of concept and application, its versatility, and the excellent results it generally produces have led to its application in all but the most stringent cases. Its greatest advantage lies in the fact that it enables the designer to

*This editorial is based on work done for the U.S. Army Research and Development Laboratories under Contract DA-36-039 SC-64625.

construct a complex filter out of basic building units called sections.** For each section an *image* impedance and an *image transfer constant* are computed from the element values of the section. (For an unsymmetrical section or half section, a different image impedance applies to each of the ports.) Most microwave engineers are more familiar with transmission-line theory than with filter theory, and for them the close analogy between lines and filters is conceptually helpful. The image impedance and transfer constant of the section are equivalent at a given frequency to the characteristic impedance and propagation constant of a length of transmission line (plus an ideal transformer if the section is unsymmetrical). Thus the image transfer constant is, in general, a complex number whose real part is the image attenuation of the section, and whose imaginary part is the image phase shift of the section. The impedance and transmission properties of a cascade of filter sections may be computed by the same equations that apply to a cascade of transmission-line lengths. In particular, if the image impedances of adjacent terminals in the cascade of sections are equal to each other at all frequencies, the total image transfer constant (and hence total image attenuation and phase) of the cascade is simply the sum of the values for the individual sections. It is this additive property that gives the image-parameter design method its great power, since the engineer has at his disposal large classes of sections with identical image-impedance functions, but with a great variety of image-attenuation functions. Thus a filter may be easily and quickly synthesized to meet almost any pass-band and stop-band insertion-loss requirements by the proper choice of filter sections. The image-impedance function of the filter sections varies with frequency, and is in general only a fair match to a constant generator or load resistance. However, by placing a particularly suited type of half section (analogous to a lumped-constant m-derived half section) at each end of the filter, a modified image-impedance function at the filter terminals is easily provided that is virtually constant over most of the pass band. Or in other cases, the pass-band impedance match may be made acceptable by means of transforming end sections. These usually are of the same type as the internal sections, but are made to be somewhat wider in bandwidth.[1†]

The image-parameter method may be applied readily to any type of filter whether low-pass, high-pass, band-pass, or band-stop. It is equally rigorous for lumped-constant and distributed-constant elements. A filter carefully designed by the image-parameter method will provide a quality of impedance match and insertion-loss performance that is very near the best that is possible in any physically realizable filter having the same number of elements.[1, 2]

Synthesis for Prescribed Insertion Loss

The image-parameter method has the shortcoming (although usually not a serious one) of approximating the most suitable insertion-loss function, rather than yielding it exactly. Other methods have been developed for synthesizing filters to have a specified insertion-loss function, if the function belongs to a class that leads to physically realizable filter networks.[2] The curves shown in Figure 2 are typical of the desirable types of insertion-loss function that satisfy this condition. The function at the top of the figure is known as the maximally flat (or Butterworth) response. It is characterized by an exceedingly low, non-fluctuating insertion loss in the pass band, increasing monotonically in the stop band. The abruptness of the change from very low to very high loss increases with the number of elements in the filter. A general formula for the element values of a lumped-constant ladder network

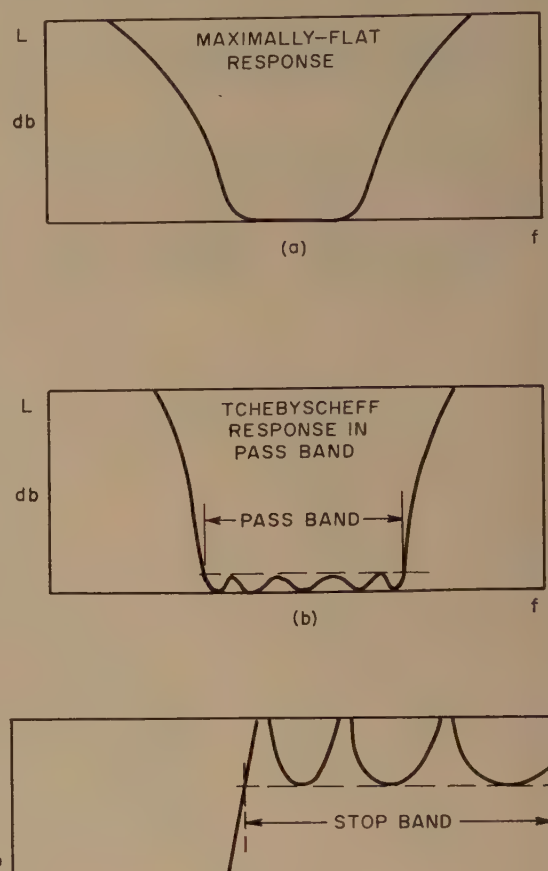


Figure 2 — Desirable insertion-loss response functions that may be obtained by filter synthesis.

to produce a low-pass response of this type was derived by Bennett, and may be applied by means of straightforward frequency-variable transformations to high-pass, band-pass, and band-rejection filters.

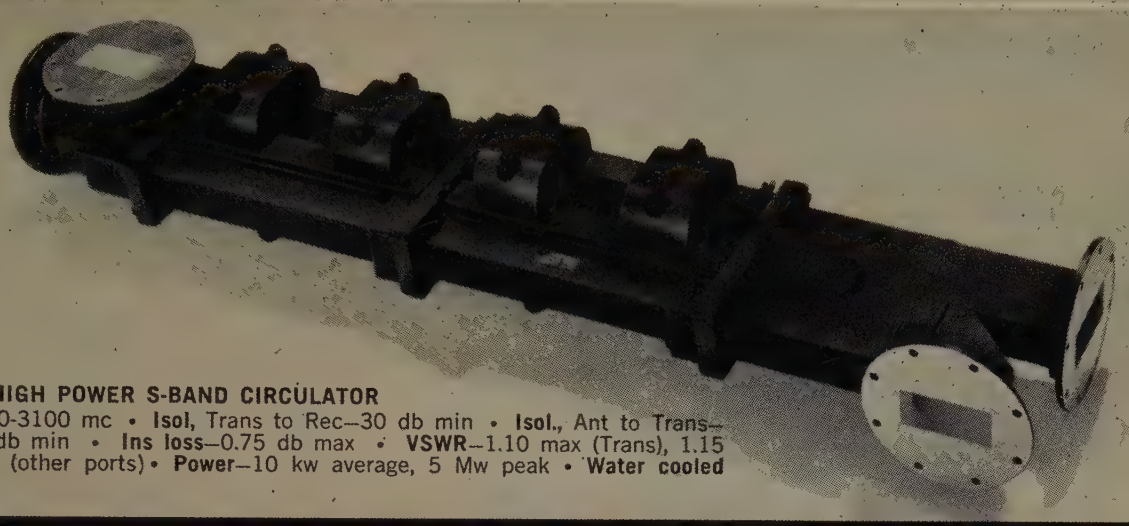
Although the smooth, nonfluctuating response of the maximally flat function is highly desirable for many applications, this pass-band behavior is obtained only with loss in steepness of cutoff in the transition region. If permitting insertion-loss ripples to occur in the pass band as in Figure 2(b), a greatly increased rate of cutoff for a given pass-band tolerance may be obtained for the same number of elements, or for a specified sharpness of cutoff and pass-band tolerance a design may be achieved with

(*Continued on page 14*)

**Even more fundamental as a filter building block is the half section, which is obtained from a symmetrical full section by bisecting it at its center.

†For the sake of brevity and readability, only a few typical references are made to the published literature. However, numerous additional papers are listed in these referenced sources. To the authors of many other excellent articles the author extends his sincere regrets and begs their indulgence.

Recent Raytheon developments in microwave ferrite devices



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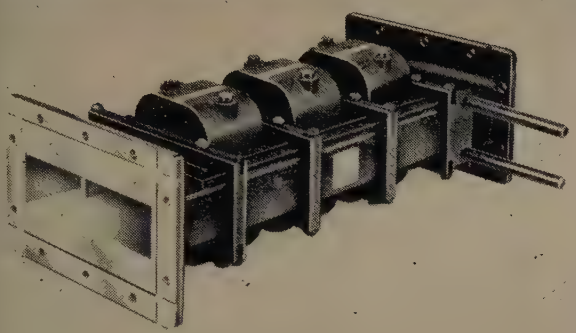

2. HIGH POWER L-BAND ISOLATOR
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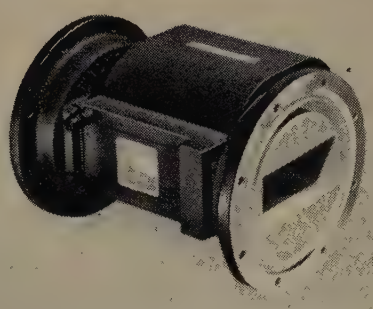

3. HIGH POWER S-BAND ISOLATOR
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substantially fewer elements. The *equal ripple* response of the figure is usually considered to be the most desirable compromise, and is known as Tchebyscheff behavior. The greater the degree of ripple, the more sharply the filter will cut off. Therefore an optimum design from the standpoint of economy of elements may be considered to be one in which the ripple amplitude is made equal to the maximum allowable pass-band insertion loss, and the number of elements is chosen so that the stop-band requirements are just achieved. A general method of synthesis for lumped-constant filters of this type was developed independently by Darlington in the United States and Piloty in Germany. A relatively simple recursion formula for the element values has more recently been derived by Belevich.

A still more desirable response from the standpoint of economy is that of Figure 2(c), in which infinite-rejection points are inserted in the stop band by means of series-resonant shunt arms or parallel-resonant series arms in the ladder network. This function is seen to have equal maxima in the pass band and equal minima in the stop band. The filter structure and its stop-band response are similar to what may be obtained with an image-parameter m-derived design, but the equal-ripple pass-band response may sometimes (although not always) lead to a somewhat more economical filter. However, any such economy will be obtained only with considerably greater difficulty in the design work. The method of network synthesis for this type of response also was developed independently by Darlington and Piloty. In fact, this method is sufficiently general to permit the synthesis operation for any response function capable of leading to a lumped-constant network consisting of nondissipative elements.

The synthesis of distributed-constant filters for prescribed insertion loss is not a straightforward extension of the lumped-constant design techniques described above. One approach has been that of deducing an approximate equivalent circuit for the distributed-constant filter, and then making the element values of this approximate circuit equal the corresponding values of a lumped-constant design having the desired insertion-loss behavior. This can be done quite easily and with very high accuracy in the case of narrow-bandwidth band-pass filters constructed of resonant cavities coupled in cascade.[3, 4, 5, 6, 7] For bandwidths greater than about 20 percent, the simpler approaches fail, and more complex relationships are necessary between the distributed-constant structure and the lumped-constant prototype. This problem is not yet fully solved, but considerable progress is being made.

Although in the general case the transcendental nature of the insertion-loss and impedance functions make exact synthesis of distributed-constant filters impossible by currently known methods, there are an increasing number of special cases for which exact synthesis has been achieved. These are all based on P. I. Richards' theorem, which permits the transcendental functions pertaining to a distributed-constant network to be converted into rational algebraic functions. For this to be possible it is necessary that all transmission-line lengths be equal (or at least commensurate) at all frequencies, and that no other elements be present in the network other than constant resistors. After the insertion-loss and impedance functions have been thus transformed, the known methods of algebra and complex-variable theory may be applied to achieve the desired synthesis.

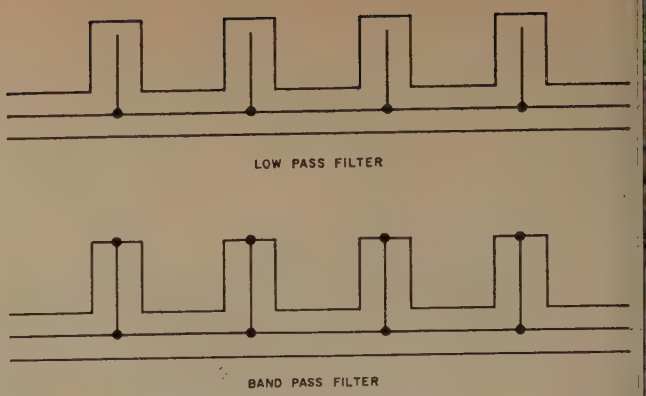

Figure 3 — Equal-line-length filters for which exact synthesis has been achieved.

In this manner, E. M. T. Jones has achieved the exact synthesis of low- and band-pass equal-ripple filters consisting of a cascade of alternate series and shunt equal-length lines, as shown in Figure 3.[8] When the ends of the shunt stubs are open circuited a low-pass response is obtained, and when short circuited the response is band pass.

Using a network transformation established by Kuroda in Japan, Ozaki and Ishii have made a large variety of transmission-line filter networks susceptible to exact synthesis.[9, 10] Since their method is based on Richards' theorem, the line lengths in all cases must be equal (or commensurate). In addition to direct junctions of line lengths, they have also treated networks containing parallel-coupled lines. A further class of transmission-line filters has been treated by an exact-synthesis method by Seidel.[11]

Conclusion

At this time, the image-parameter method is usually favored over exact synthesis methods for the design of low-pass and wide-bandwidth band-pass microwave filters. This is true because the image-parameter method usually results in performance nearly equal to the best theoretically possible, and with a small fraction of the labor of an exact synthesis design. The image-parameter method also permits transmission lines of any desired lengths to be used in the filter network, either by themselves or in combination with lumped capacitive and inductive elements. This enables one easily to place infinite rejection points wherever needed in the stop-band region, whereas with exact synthesis methods, this is much more difficult. It also permits the use of a far larger number of practical structural configurations, with discontinuity effects taken into account.

An interesting feature of transmission-line filters designed by exact synthesis is their periodic frequency response. In many applications this may be a disadvantage from the spurious-response standpoint. Filters designed by image-parameter or approximate-synthesis methods are not necessarily restricted in this way.

For applications where the response functions and configurations of exact-synthesis filters are suited, that kind of design will of course be favored, especially once numerical design data become available. In other cases, one or another of various approximate synthesis techniques now under study may prove to have special advantages.

With narrow-bandwidth band-pass filters having bandwidths up to about 20 percent, the available approximate synthesis methods for coupled-resonator filters are simple

(*Continued on page 20*)

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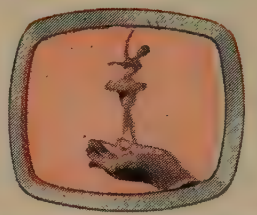

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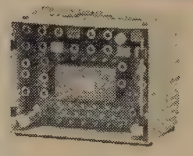

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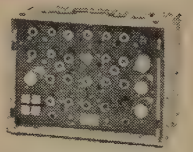

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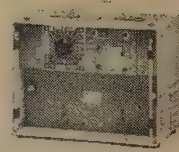

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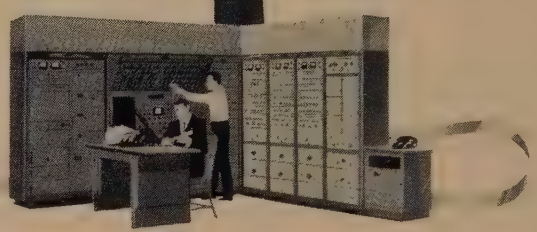

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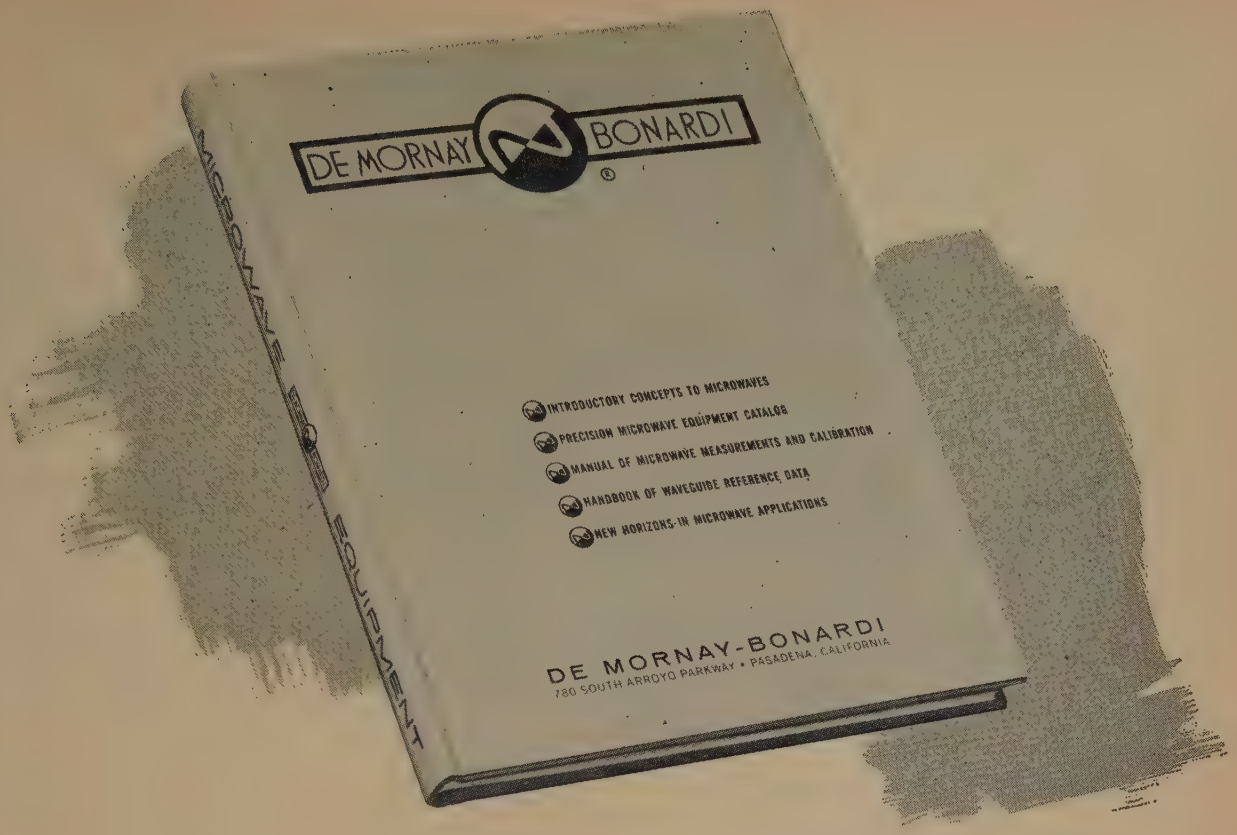
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(*Continued from page 1*

and accurate. They do not provide i finite-rejection points in the stop ba however, which can be a handic when very sharp cut off is desired.

To summarize, almost any desir filter response (if theoretically fe ible) can be achieved in practice to good approximation by one or a other of the currently available desi methods and structural configuratio However, future research will doul less make possible the design of given filter with less effort, greater a curacy, and a simpler, more compa structure than is possible today.

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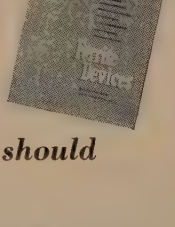

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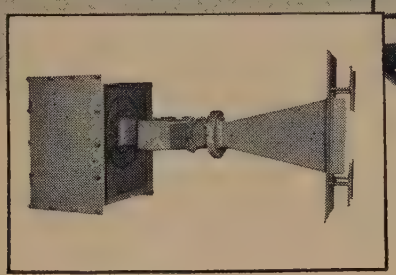

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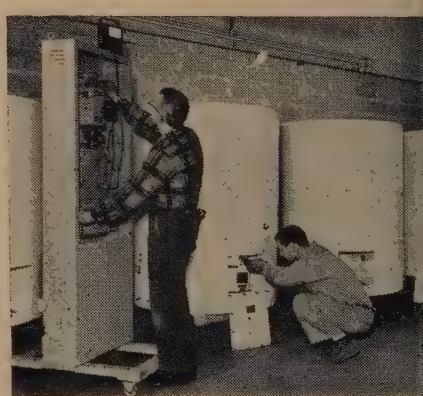

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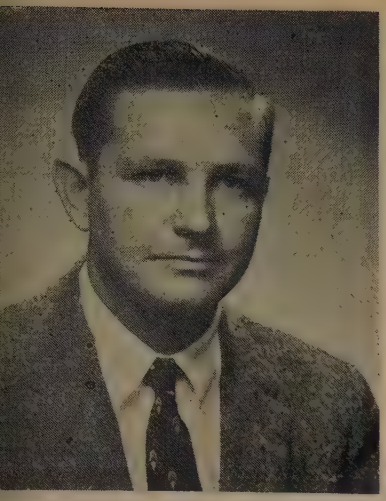

JOSEPH P. SWANSON, Assistant Chief Engineer. EE and MS(EE) Stanford University, BS(EE) Antioch College. As a member of the system study group of Electronic Research Laboratory of Stanford University, Mr. Swanson worked on countermeasures applications of traveling-wave tubes and associated devices, and supervised construction of broad-band repeater systems and search receivers. At Levinthal, he has been in charge of numerous projects including modulators for modulating-anode and gridded klystrons, for pulse-group klystron operation, and for shaped pulse klystron operation.

P. E. TANNENWALD was born in Kiel, Germany in 1926, came to the United States in 1940 and became a U.S. citizen in 1945 while serving in the U.S. Army. He did his undergraduate as well as graduate work at the University of California, Berkeley, receiving the Ph.D. degree in nuclear physics in 1952. Dr. Tannenwald worked at the UC Radiation Laboratory in high-energy particle physics before accepting his present position with the Massachusetts Institute of Technology. At Lincoln Laboratory he has been engaged in the microwave aspects of solid state research —in such areas as ferromagnetic resonance in ferrites and garnets, and masers. It was the initiation of microwave resonance experiments in thin magnetic films that led, in collaboration with M. H. Seavey, Jr., to the direct observation of spin wave resonance.

Dr. Tannenwald is a member of the American Physical Society, Phi Beta Kappa and Sigma Xi, and is at the present time Assistant Group Leader of the Resonance Physics Group at Lincoln Laboratory.

INDUSTRY NEWS

CASWELL ELECTRONICS — A NEW COMPANY

Announcement has been made of the organization of the Caswell Electronics Corporation, which will specialize in the development and production of microwave transmission line components and subassemblies. The plant is located at 1414 Queens Lane, San Jose, California.

Mr. Dwight Caswell, President of the new corporation and formerly Director of Research and co-founder of Cascade Research Corporation, announced that employment is expected to reach thirty within a year.

The company is now designing several unique ferrite-containing microwave devices. Design emphasis is being placed on coaxial line techniques and new types of waveguide components. The company is seeking to fill the need for a quick-acting combined engineering and production facility.

• • •

KANSAS MICROWAVE RELAY SYSTEM EXTENDED

A three-station addition to Kansas Highway Commission's existing microwave relay system is now being installed by General Electric's Communication Products Department.

The new extension will provide communications from the commission's control center at Norton, in northwestern Kansas, near the Nebraska border, southeastward to Stockton.

The Kansas Highway Commission placed its first microwave equipment in operation last summer as a means of centralizing its radio control. At that time, four stations were installed by General Electric west of Norton, terminating at Colby.

The number of circuits desired by the Commission for radio communications led Kansas Highway officials to choose microwave for its complex repeater operation.

• • •

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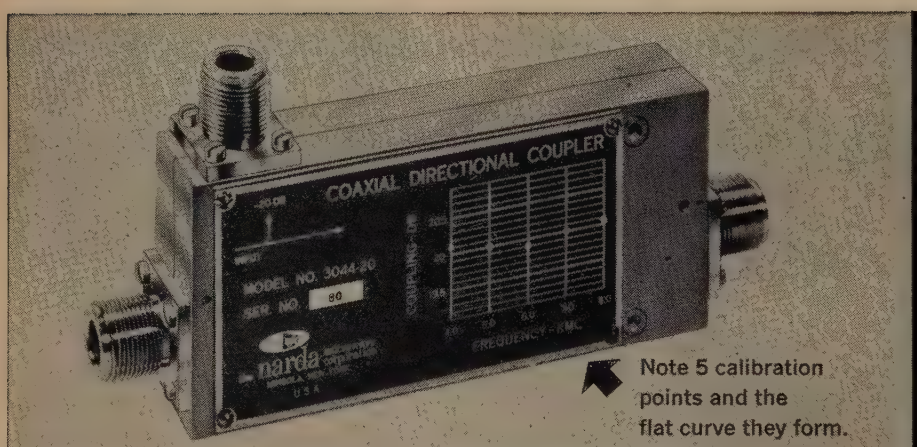

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2000-4000	20	3042-20	1.15/1.2	20	1000	200	10	
4000-8000	20	3044-20	1.2/1.25	17	1000	200	10	
7000-11,000	20	3045-20	1.25/1.3	15	1000	200	10	

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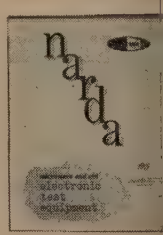

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SPIN WAVES*

P. E. Tannenwald

M. I. T. Lincoln Laboratory

Introduction

There are two reasons why the relatively esoteric concept of spin waves has now entered the province of the microwave engineer. The first contact came when considerations more subtle than classical physics had to be invoked to understand high-power phenomena in ferrite devices and describe certain modes of operation of the ferromagnetic amplifier. The second reason is quite general: if the microwave engineer is interested in utilizing the interaction of microwaves with ferromagnetic media, he will come to realize that spin waves are the appropriate natural modes of propagation in a system of coupled spins. Electromagnetic waves are natural modes, too, but they utilize different properties of the medium. Just as was the case with masers, the basic processes are quantum mechanical in nature; however, a qualitative understanding can be obtained in terms of a classical description. This is the approach we shall use then — a descriptive, almost pictorial, representation of spin waves, which, for a first acquaintance with the subject seems justifiable.

Spin wave motion becomes quite a reasonable concept if we look back at other types of wave motion with which we are already familiar. For example, elastic waves or sound waves propagate in those media which can be distorted longitudinally and which have restoring forces of magnitudes such as to give rise to the appropriate frequency, wavelength and velocity. Electromagnetic waves are vibrations in either the fictitious medium called "ether" or in real media with appropriate atomic properties. The analogy thus gives the clue: sound waves can propagate in media consisting of molecules and having inter-molecular forces or bonds, electromagnetic waves can propagate in media where the electric and magnetic fields can interact with electronic charges and moments, while spin waves can propagate in ferromagnetic media consisting of spins and having interspin forces. Thus spin waves are the natural modes of vibration or propagation of a system of interacting ferromagnetic spins. The inter-spin forces are the "exchange forces" which give rise to the phenomenon of ferromagnetism in the first place.

*The work reported in this paper was performed by Lincoln Laboratory, a center for research operated by Massachusetts Institute of Technology with the joint support of the U.S. Army, Navy, and Air Force.

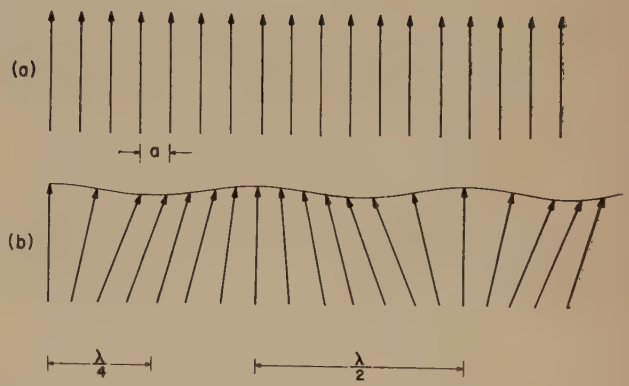

Figure 1 (a) — One-dimensional chain of spins.
(b) — Plane-vibrational spin wave.

We are already in a position to distinguish between electromagnetic wave and spin wave propagation in ferromagnetic media. Ordinary electromagnetic wave motion in a medium usually involves displacement of electric charges and magnetic moments against electric forces — as described by the macroscopic parameters of dielectric constant and permeability; spin wave motion occurs whenever spins are tipped against the exchange forces. The exchange field which holds spins in alignment in ferromagnetic materials is well-known to be of exceedingly high magnitude compared to ordinary laboratory fields — say, 1 million gauss. Consequently we might expect a *priori* that the frequency of vibration resulting from displacement against such a restoring force will be high, that excitation might be difficult or at least take place only under unusual circumstances, and that detection might have to be indirect.

Pictorial Representation

Let us now look into the interior of a ferromagnetic medium and idealize the structure by a row or chain of parallel spins. (Figure 1a) A propagating spin wave is then depicted simply (and somewhat incorrectly) as a traveling wave disturbance of one spin after another successively, in time and space, tipping aside and returning to equilibrium (Figure 1b). A standing spin wave would consist of different spins at different points in space vibrating with different amplitudes.

In order to derive the vibration frequency of a spin wave (or, more precisely, the dispersion law), we first find an expression for the energy contained in a spin wave. If equilibrium exists when all spins are lined up, then it is reasonable to expect that the energy between two spins making a small angle ϕ with each other should be proportional to ϕ^2. In particular, $E = JS^2\phi^2$, where E is the "exchange energy," J is a proportionality constant which expresses the strength of exchange coupling between spins, and S is the magnitude of a spin. Figure 2 shows the case of maximum displacement ($\phi_{max.}$) to one side, which evidently occurs every quarter wavelength. Furthermore, if the distance between individual spins is a, then the angle ϕ between neighboring spins is approximately** $\phi = \phi_{max.} \frac{2\pi a}{\lambda}$. Converting λ into the wave vector k, we obtain for the energy of N spins, $E = NJS^2\ \phi^2_{max.} k^2a^2$.

In the next step we try to relate $\phi_{max.}$ to some externally observable quantity. The saturation magnetization is clearly proportional to NS, so that when spin wave motion takes place and spins tip away from the lined-up position, the net magnetization will decrease. Such a decrease can be expressed in terms of the ϵ shown in Figure 2. Further, from geometry, $\phi^2 = \frac{2\epsilon}{S}$; so now

$$E = NJS^2 \frac{2\epsilon_{max.}}{S} k^2a^2 = 2JS\, k^2a^2 \cdot N\epsilon_{max.}$$

For the final step we must resort to a simple quantum-mechanical argument. If a spin wave contains an energy E, it can be excited by n photons containing an equal amount of energy. Thus

$$n\frac{h\omega}{2\pi} = 2JS\ k^2a^2 \cdot N\epsilon_{max.} \qquad (1)$$

This equation gives us our principal result — namely the dispersion relation for spin waves, $\omega \propto k^2$. (This differs from the dispersion relation for electromagnetic waves, where $k^2 = \mu\epsilon\omega^2$.) We also know from quantum mechanics that a given photon cannot tilt a spin by some arbitrary amount, as shown in the figures by a classical model. Rather, it either flips a spin S (if $S = \frac{1}{2}$) completely over, or it does nothing at all. Consequently, the quantum mechanical interpretation of Equation 1 is that the number of complete spin reversals must be $n = \frac{N\epsilon_{max.}}{S}$. This is physically quite reasonable because the same result is obvious from simply looking at the figures. The decrease of net magnetization is evidently some fraction of the original total spin amplitude NS, namely $NS \cdot \frac{\epsilon_{max.}}{S} = N\epsilon_{max.}$. What the classical picture does then is to take the quantum-mechanical, complete spin reversals of a few spins and distribute them equally over all the spins so as to create a continuous wave-like disturbance.

At this point we may well ask what useful purposes the above spin wave model can serve. The fact is that, although spin waves had not been observed directly until very recently, postulating their existence some 30 years ago has led to a number of explanations of magnetic behavior. For example, the decrease of saturation magnetization with increasing temperature and the final disappearance of magnetism above the Curie point, can be viewed as the result of thermal agitation of the spins. A quantitative law can be derived for the magnetization as f(T) if it is assumed that the amplitude of spin wave excitation increases with increasing temperature.

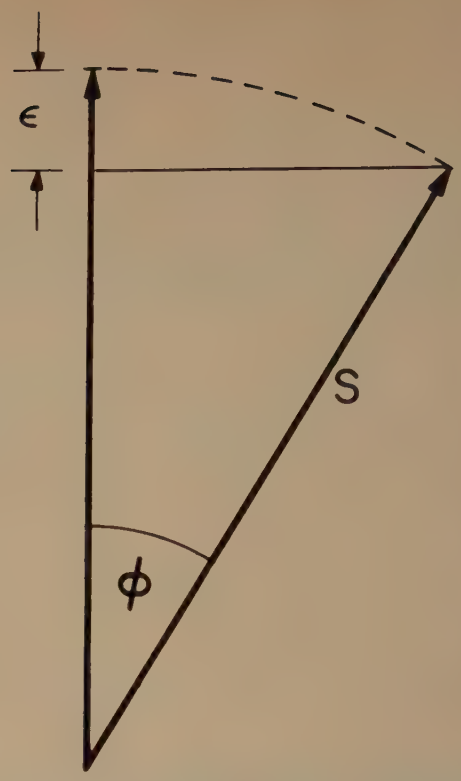

Figure 2 — Geometry for showing that excitation of a spin wave diminishes net spin by amount ϵ.

Spin Waves in Ferromagnetic Resonance

So far we have committed one gross oversimplification in ignoring the fact that the actual response of a spin to displacement from equilibrium is not plane-vibrational but rather precessional. We should now extend our concept of spin waves and visualize the wave motion occurring superposed on the precession. (See Figure 3 a, b, c). If only thermal agitation is present, both the precession and the spin wave motion will be random and incoherent; therefore neither one is detectable externally. At least as far as precession is concerned, we know how to make it externally detectable — we impose an r.f. driving field to supply coherent energy. With the addition of an external DC field to tune the frequency into the microwave range, these are just the conditions for ferromagnetic resonance (FMR). *A priori* there exists no good reason why the spins should not all precess together in phase and with the same amplitude. Thus all spins are parallel at all times, no exchange forces are called into play, and we have no spin waves. However for elegance's sake, the uniform precession mode of FMR may be called a spin wave of infinite wavelength or $k = 0$ wave number.

It now becomes apparent that some unusual condition will have to prevail before spins can be disengaged from their concerted parallel motion. A first attempt might be to place the magnetic sample into an inhomogeneous r.f. field, let us say in certain regions in a microwave cavity. This type of experiment has been carried out and indeed gives rise to non-uniform precession. However the types of r.f. field gradients available in this manner are far too

**For simplicity, correct numerical averaging factors have been omitted in what follows.

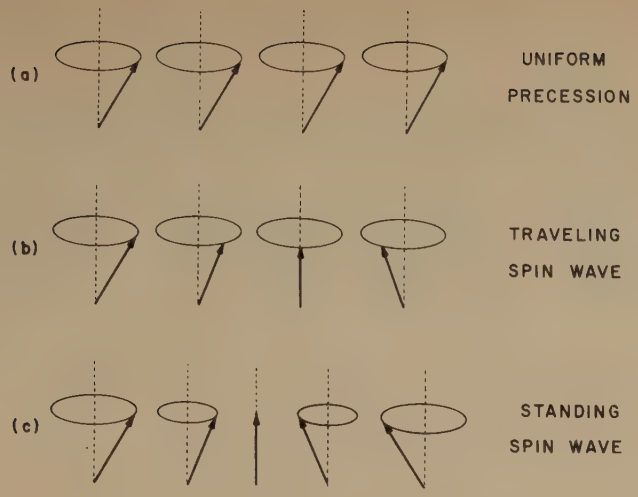

Figure 3 (a) — Uniform precession of spins.
(b) — Traveling spin wave.
(c) — Standing spin wave.

small to cause anything but a negligible tipping between adjacent spins. The non-uniform modes arising in this manner are called magnetostatic, are essentially electromagnetic in character, and do not involve exchange forces. What is needed is a naturally existing gradient on an atomic scale. Such a gradient is ready-made for us within a skin depth layer when an r.f. field penetrates a good conductor. (See Figure 4). It can be shown that under such circumstances there does exist a spin wave contribution to the FMR phenomenon. In effect, a $\frac{\lambda}{4}$ standing spin wave is set up between the surface of the metal and an interior layer where the r.f. field is practically zero.

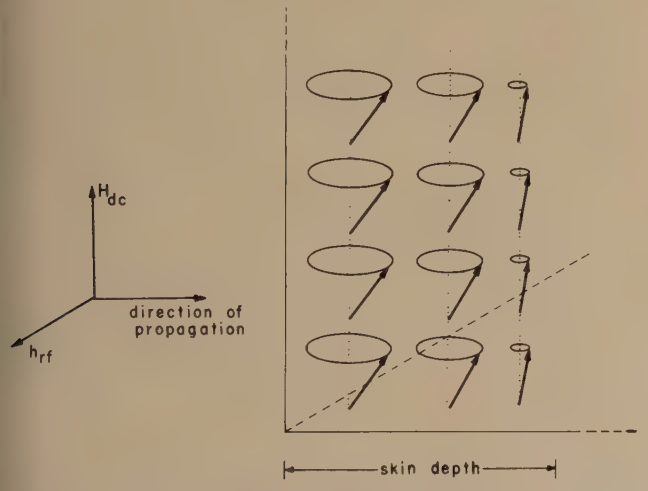

Figure 4 — Pseudo standing spin wave excited due to skin depth phenomenon.

Another suitable condition might be created if it were possible to select some arbitrary spin, hold it fast, and drive the rest of the spins into excitation. The mechanical analogue of this situation is a string fixed at one or both ends and driven into vibration at various harmonics. For the spin system such a scheme can be devised by going to an exceedingly thin slab or film of magnetic material so that dimensions only a few orders of magnitude larger than atomic come into play. It turns out, for reasons not completely understood, that the end spins at the surface boundaries find themselves in sufficiently different environment (due to lack of symmetrical neighbors) to behave as if they were fixed. Under such favorable circumstances it has indeed become possible for an applied r.f. field to excite higher order modes — that is, spin wave modes. The situation within the thin layer is shown schematically in Figure 5. The actual experiment was carried out in an evaporated film of permalloy mounted in a microwave cavity which was part of a standard FMR setup. The striking feature of the resulting absorption curve, shown in Figure 6, is the well-defined system of harmonics, which have been identified as spin wave vibrations of different wavelengths. Using the applied r.f. field as a constant frequency drive and the external DC field for tuning purposes, the sample is then seen to absorb energy whenever the condition is satisfied that an integral number of half-wavelengths fit into the film thickness L. A demonstration of this kind — called spin wave resonance — helps in understanding more complicated phenomena involving spin waves in ferromagnetic resonance.

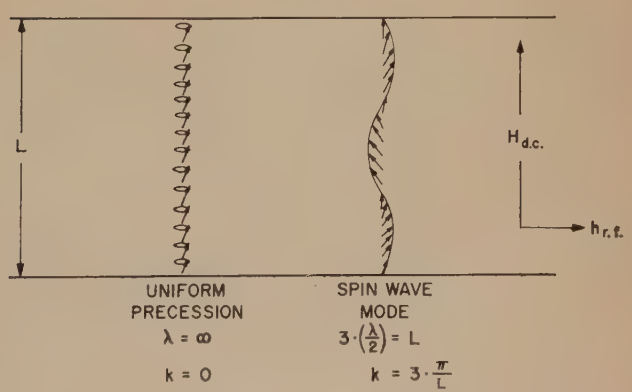

Figure 5 — Pictorial representation of spin wave resonance in a thin film.

Applications of the Spin Wave Concept

The types of spin wave experiments described above are of obvious physical interest because they provide access to measurement of the basic ferromagnetic exchange inter-

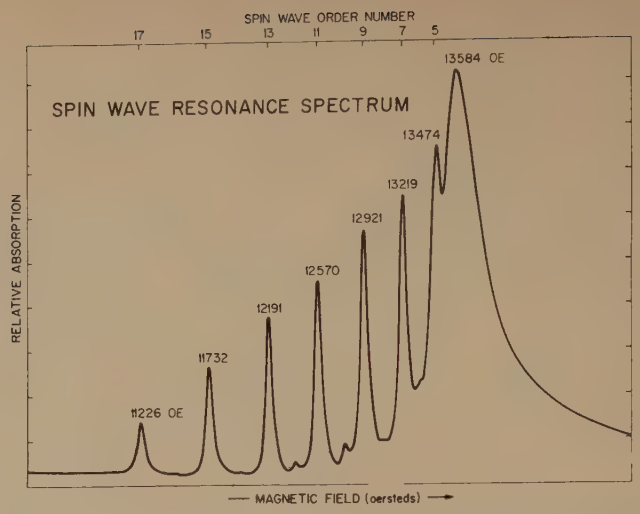

Figure 6 — Microwave absorption curve showing well-defined resonance peaks whenever interference condition for spin wave harmonics is satisfied. The sample is a 3900 Å film of permalloy in a microwave cavity.

action. But perhaps more important to the microwave engineer, the spin wave concept allows for interpretations of some otherwise unexplained FMR phenomena. For example, the width of the resonance line in FMR is thought to result from the transfer of r.f. energy in the uniform precession to large numbers of spin waves which happen to have the same frequency but a whole variety of possible wavelengths. The particular spin waves to which energy is transferred are determined by boundary conditions imposed by grains or local inhomogeneities.

Another FMR phenomenon which can be explained in terms of the spin wave concept is the instabilities which occur under high microwave power. It can be shown that at a certain critical r.f. field level, a positive feedback condition develops in which spin waves are excited until a drastically different equilibrium situation prevails. Furthermore, it has been proposed that spin wave modes can be used in the ferromagnetic type of parametric amplifier, where the precession provides the nonlinear coupling between modes.

Conclusion

In concluding our description of spin waves, we may take the liberty of indulging in a bit of speculation. We have tried to emphasize that spin waves are the natural type of wave motion in materials consisting of a system of coupled spins, just as sound waves are in media consisting of coupled molecules. Then we may ask, is it possible for spin waves to carry information? The answer is yes except for two difficulties. First, in all presently known media spin wave motion is subject to very high attenuation, just as precessional motion is. Second, since the dispersion law is of the form $\omega \propto k^2$, the group velocity, which carries the energy in a wave packet, is proportional to k and thus subject to dispersion. Actually the second difficulty is absent in antiferromagnetic materials — a special class of magnetic media in which adjacent spins are anti-parallel to each other. For an anti-ferromagnetic chain of spins we could carry out a similar analysis as above, and we would find that the dispersion law is $\omega \propto k$.

Even though we have discussed electromagnetic and spin waves as separate entities, they will always appear physically mixed together because the only presently known coherent excitation means is electromagnetic. Ultimately, it is conceivable that there may be sufficient coupling between spin waves and lattice vibrations so that spin waves can excite ultra-high frequency sound waves or vice versa. Thus, as seems so frequently to be the pattern, scientific advances answer some questions, raise many more new ones, and also provide a glimpse of exciting things to come.

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MODULATOR TECHNIQUES FOR GRIDDED KLYSTRONS AND TRAVELING WAVE TUBES

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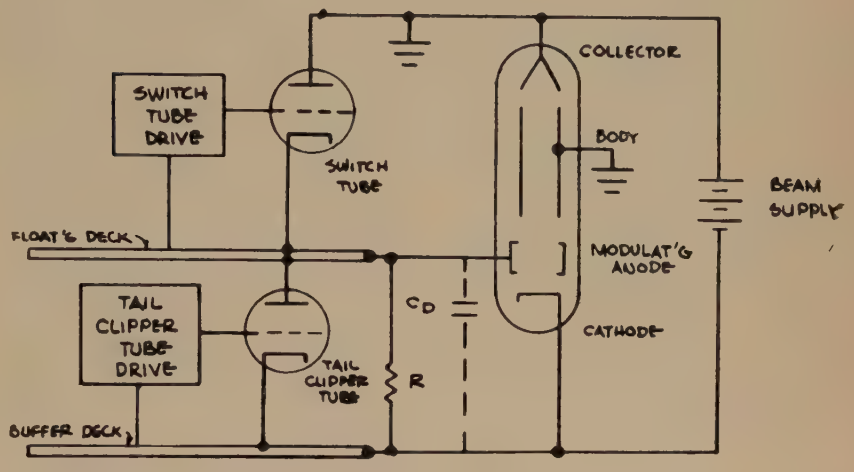
Introduction of control elements in some of the high-power klystrons and traveling-wave tubes now being built by such firms as Eitel-McCullough, Varian Associates, Sperry Gyroscope, and others, has called for development of new types of modulator elements designated as "floating-deck" circuits. Preliminary to a description of these circuits, a summary of the properties of gridded klystrons and traveling-wave tubes as load elements may be useful.

Generally, in high-power tubes, the control grid, or modulating anode, as it is variously called, is a ring which intercepts only a very small fraction of the total beam current. Consequently, the gun perveance is low and the amount of voltage which must be applied to cause full beam current is on the order of 30% to 100% of full beam voltage.

As a load, this control element may be represented as a high-impedance resistor (1 megohm or more) shunted by a capacitance which may vary between 25 and 100 $\mu\mu$f. Modulators for gridded klystrons and traveling-wave tubes are therefore characterized by the requirement that they must deliver a large positive-going voltage to a load that is essentially capacitive, and the fact that the modulated beam current need not pass through the modulator.

Figure 1 shows the basic floating-deck modulator circuit in simplified form. The tube to be modulated is connected directly across the main beam supply. The modulator consists essentially of two high-voltage tubes connected in series. The uppermost of these tubes, designated as the switch tube, has its plate connected to ground and its cathode connected to the chassis designated as the floating deck.



Pulser Properties

Leading edge — linear rise $\frac{dv}{dt} = \frac{I_s}{C_d}$

Trailing edge — linear fall $\frac{dv}{dt} = \frac{I_s}{C_d}$

Pulses per second — limited by plate dissipation $f = \frac{2P}{cv^2}$

Pulse width — unlimited
Pulse duty cycle — unlimited
Pulse sag — unlimited
} except for Beam Supply

v = pulse voltage

I_s = peak switching current

P = plate dissipation of switch tube

f = pulses per second

C_d = total system capacity—floating deck and load capacity

Figure 1 — The basic Floating Deck Pulser.

The second high-voltage tube acts as a tail clipper and has its cathode connected to the negative terminal of the beam supply and its anode connected to the floating deck. The klystron modulating anode is also connected to the floating deck.

Operation of the basic floating-deck pulser circuit may be described as follows: During the interpulse period, both the switch tube and the tail clipper tube are non-conducting, and the klystron modulating anode is held at cathode potential by means of a high-impedance bleeder resistor, R. The leading edge of the modulator output pulse is formed when the switch-tube drive circuit pulses the switch tube into conduction and shorts the floating deck and hence the klystron modulating anode to ground potential.

During the flat-top portion of the pulse, the switch tube is held on in the bottomed condition, thus clamping the modulating anode to ground. The trailing edge of the pulse is formed when the tail clipper tube is pulsed to conduction by the tail-clipper drive circuit. At this time, the switch-tube drive circuit is turned off and the

tail-clipper tube shorts the floating deck back to the beam-supply negative potential.

A circuit of this type has several interesting properties: First, it will be noted that the pulse leading edge rises linearly with time in accordance with the relation $dv/dt = I_s/C_d$, and that the trailing edge of the pulse falls linearly in accordance with the same relation. Furthermore, the speed of the pulse leading and trailing edge is completely independent of the pulse width. Because of the nature of this circuit, pulse overshoot should be negligible, sag during the flat top of the pulse should not be present, and the pulse generator should be subject to repetition-rate but not duty-cycle limitations.

The basic limitations which do exist in this circuit, stem from limitations of currently available switch tubes. First, the switch tube must obviously stand off the full beam voltage. Second, leading-edge rate of rise is limited by the combination of available peak switch current, I_s, and total load capacitance, C_d. Finally, the total number of pulses per second which can be delivered by the modulator is limited by the allowable plate dissipation of the switch tube.

This may be seen as follows: With each pulse an amount of energy equal to $C_dv^2/2$ is stored on the capacitor C_d, and a like amount of energy is dissipated in the switch-tube plate. Thus switch-tube plate dissipation, P, may be expressed as $C_dv^2f/2$ and the limiting number of pulses per second is expressed as $F = 2P/C_dv^2$.

It should be noted that this repetition-rate limitation is on a total pulses-per-second basis. Much higher pulse rates could be handled on a pulse-group basis where the group duration is short compared to the switch-tube thermal time constant.

Since the basic limitations of floating-deck pulsers are determined completely by the properties of available switch tubes, it is of interest to show the range of specifications which can be attained with several types of currently available tubes. This has been done in Figure 2. The straight sloping line relates the 0 to 100% pulse rise time (right-hand scale) to the required pulse voltage, and the curved line denotes the maximum pulse-repetition rate which can be obtained at this pulse voltage.

Thus, for example, one sees that a 20-kv floating-deck pulser using 4-1000A tubes as both the switch tube and tail clipper tube would have a maximum repetition rate of 25 kc and a 0.27 μsec rise and fall time; or that X-556D switch tubes can be used for 50-kv pulses at frequencies up to 2 kc with rise times just over 1 μsec.

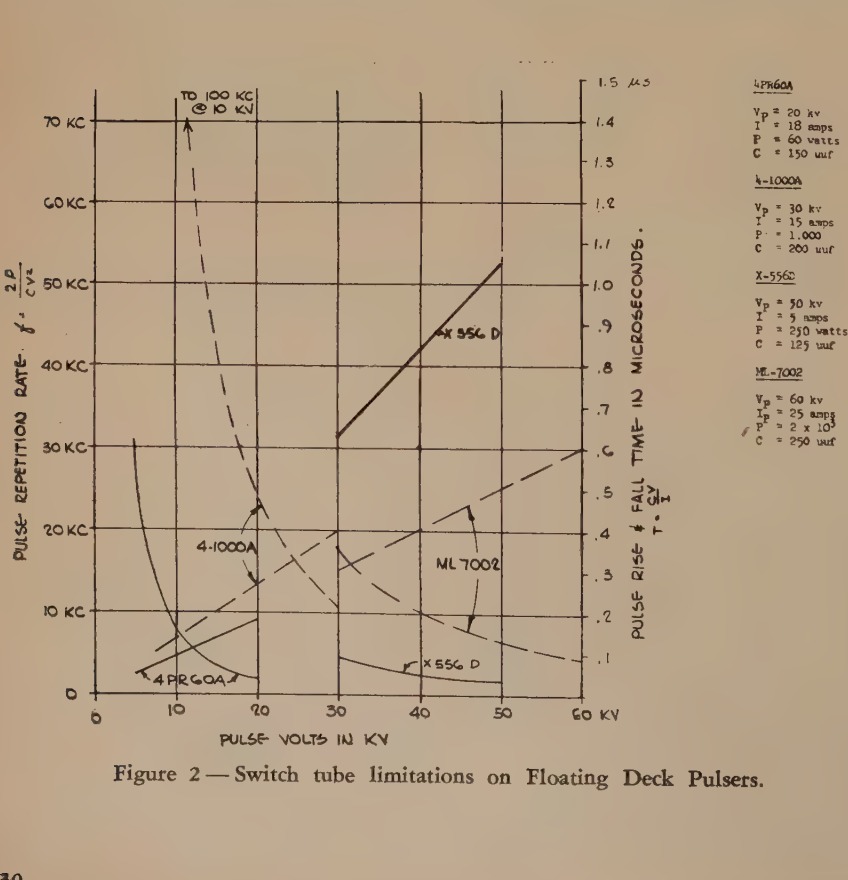


Figure 2 — Switch tube limitations on Floating Deck Pulsers.

Two additional facts should be noted. First, different values of load capacitance C_d have been used with different tube types. These variations represent an attempt to estimate the total capacitance likely to be encountered when using these tubes under typical conditions. Finally, paralleling two switch tubes will not quite double the pulser performance indicated on this graph since the second tube will undoubtedly introduce additional load capacitance.

Three Examples

The foregoing has assumed ideal drive circuits for both switch tubes. Except for the case of pulses with very short rise times this assumption is justified, though the requirement of flexibility certainly increases drive circuit complexity.

One may illustrate this by briefly describing three different floating-deck modulator circuits: The first of these is the modulator used in the first prototype of the Sentinel radar-system transmitter. This circuit is probably as simple a floating-deck modulator as one can make. In the second circuit to be described, simplicity has been sacrificed in the interests of flexibility. The flexibility attained has been such that with only very minor changes the same modulator circuit has been used for the Columbia University ORDIL system, two MIT Lincoln Laboratory transmitters used for propagation research, a Signal Corps transmitter, two Stanford Research Institute transmitters used for auroral and meteor-echo studies, and a second version of the Sentinel radar system modulator. The final pulser circuit to be described permits bypassing some of the switch-tube limitations shown in Figure 1.

Figure 3 shows a simplified circuit diagram of the 50-kv pulse generator used in the first prototype of the Sentinel radar system. This circuit was required to produce 50-kv pulses at a repetition rate of 500 pulses per second. Two fixed pulse widths were used, 5 μsec, and 40 μsec, and rise and fall times of these pulses were 1 and 3 usec, respectively. Peak r-f power out of the transmitter was 200 kw.

Operation of this circuit may be described as follows: In the interpulse period, the floating deck is held at negative beam potential by means of the 5-megohm bleeder resistor, and the trigger-tube and tail-clipper-tube coupling capacitors are each charged to approximately 55,000 volts. The

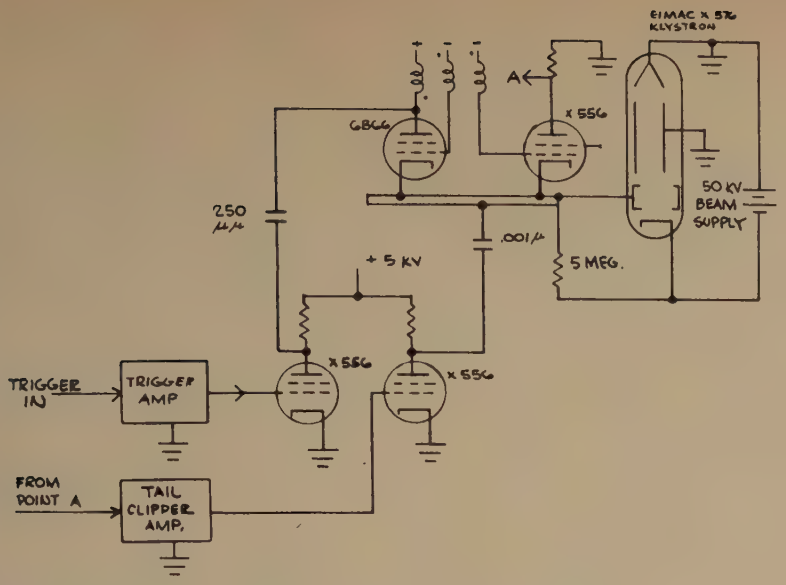


Figure 3—Simplified diagram of simple 50 KV Pulser used on Sentinel Radar prototype.

leading edge of the pulse is initiated when the trigger amplifier pulses the trigger X-556 into conduction for approximately 0.2 μsec. Conduction of this tube plate triggers the 6BG6 blocking oscillator which forms the switch-tube drive circuit, and thus causes the switch tube to short the floating deck to ground potential for the duration of the blocking-oscillator pulse.

The turning off of the switch-tube blocking-oscillator drive is detected by means of the cessation of switch-tube current as detected at point A. This signal is amplified and used to turn on the X-556 tail-clipper tube whose plate voltage has been raised to 55 kv during the modulator output pulse. Variable pulse width is obtained by means of a switch on the floating deck which changes blocking-oscillator transformers.

The simplicity of this circuit is obtained only at the expense of several severe limitations: First, the maximum pulse width for which effective tail clipping can be obtained is primarily determined by the time constant of the tail-clipper plate-load resistor (5 meg) and the tail-clipper coupling capacitor. A maximum duty cycle and maximum repetition rate are also determined by these two elements in conjunction with the 5-megohm bleeder resistor. Finally, the pulse length is not continously variable from ground level.

In conclusion, it is evident that the major limitations of this system stem from the fact that the tail-clipper tube is capacitively coupled to the floating deck.

The block diagram of a circuit which would eliminate these disadvantages is shown in Figure 4. Here both the trigger tube and the tail-clipper tube are essentially direct coupled to the floating deck and have their cathodes at —50 kv. The leading edge of the modulator pulse is formed when an "on" trigger pulses the trigger tube into conduction for approximately 0.1 μsec. Conduction of the trigger tube causes the switch-tube drive circuit to be turned on by means of the box labeled "Switch On."

This circuit is such that the switch tube will remain in a conducting condition until its drive circuit is turned off by the box labeled "Switch Off." The trailing edge of the pulse is keyed by means of a trigger signal derived from a variable time delay located at ground level. The signal is amplified in the tail-clipper-tube drive circuit and pulses the tail-clipper tube into conduction. In addition to shorting the floating deck to negative beam potential, the tail-clipper plate current is also used to trigger the "Switch Off" box and turn off the switch-tube drive circuit. Thus, both repetition rate and pulse width are determined at ground level and the problems stemming from capacitive coupling have been eliminated.

Before leaving the subject of pulse generators, there is one other approach to this problem which should be mentioned. Figure 5 shows a simplified schematic of a 30-kv pulse generator. This circuit was designed to provide 30-kv pulses to the grid of a developmental Varian multi-megawatt klystron. Beam voltage for this klystron is 120 kv. Consequently, the klystron cathode as well as the entire grid modulator circuitry was immersed in oil.

The modulator was required to provide a pulse whose width was continuously variable from 3 to 7 μsec and whose rise and fall times did not exceed 0.25 μsec. Pulse repetition rate, pulse amplitude, and pulse width were to be controlled from ground level. Operation of this circuit may

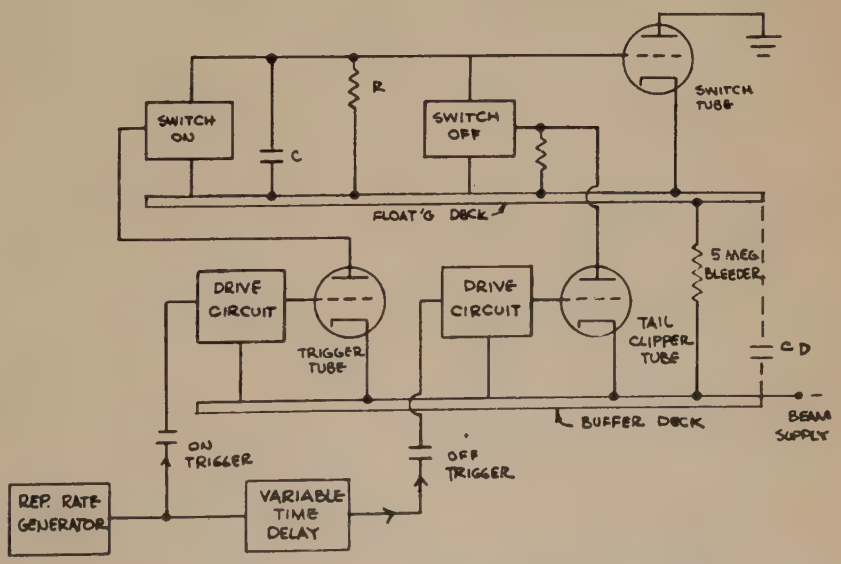


Figure 4 —Block diagram of versatile Floating Deck Pulser.

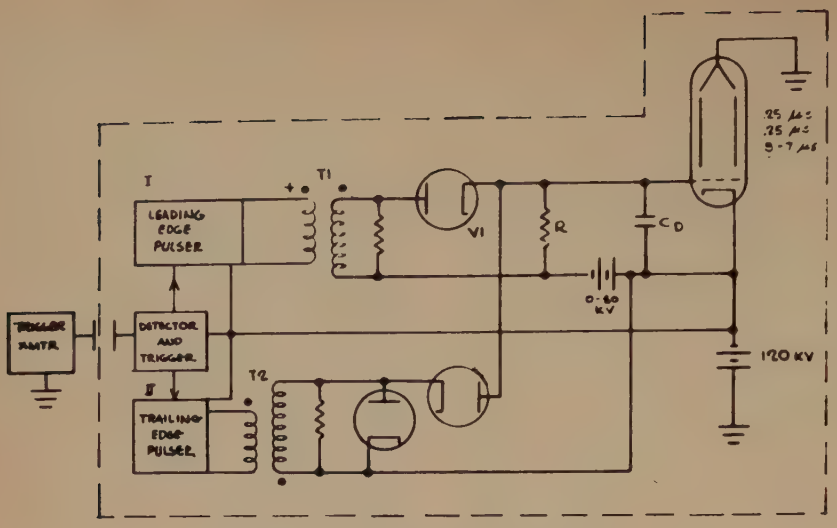


Figure 5 — The double pulse circuit.
Note: All components within dashed line in oil at — 120 KV.

be described as follows: In the interpulse period, the klystron grid is biased to —30 kv by means of a high-impedance resistor R. Plate-to-cathode voltage on diode V1 equals zero, and the plate of diode V2 is 30 kv negative with respect to its cathode.

A trigger pulse whose repetition rate and pulse width equal those of the desired modulator output pulse is transmitted by means of an r-f link to the modulator, detector, and trigger circuit floating at —120 kv. The leading edge of this trigger pulse is used to trigger a leading edge pulser. This pulser generates a short (0.5 μsec) +30 kv pulse voltage on the plate of diode V1, thus capacitor C_d, the load capacity, is charged to +30 kv. When the positive voltage is removed from the plate of V1, trapped charge on C_d maintains an approximately constant drive voltage on the grid of the klystron.

It should be noted that as the 30-kv drive signal was applied to the klystron grid it was also applied to the plate of diode V2, thus the plate-to-cathode voltage of diode V2 is zero during the flat top of the pulse. The trailing edge of the modulator output is initiated when the detector and trigger circuit cause the trailing edge pulser to apply a ½-μsec —30 kv pulse to the cathode of diode V2, and thus remove the trapped charge from capacitor C_d.

This circuit, when examined in detail, is roughly as complex as a typical floating deck pulser but by no means as flexible. It has, however, two properties which in certain circumstances enable it to meet pulse specifications which could not be met as readily by the standard floating-deck circuit. The first of these is that one may take advantage of the voltage step-up properties of pulse transformers to match the desired switch-tube operating points to the desired pulse amplitude without running into the limitation of rise-time to pulse-width ratio imposed by pulse transformers. Secondly, the maximum pulse voltage which can be obtained from this circuit is now limited by the peak inverse rating of available diodes rather than triodes or tetrodes.

Before closing, a few comments might be made about the use of floating-deck circuits in the generation of shaped rather than rectangular pulses.

The basic floating-deck modulator cir[cuit] cuit as shown in Figure 1; namely, two switch tubes back to back with load capacitance at their junction, i[s] particularly adaptable to the genera[tion] tion of shaped pulses such as thos[e] required by the Cambridge Electron Accelerator Project 6-Bev synchrotron.

Referring to the circuit of Figure 1[,] one sees that at any time, t, the voltage on the klystron modulating anode V_k equals the integral of the switch-tube current minus the integral of the tail clipper-tube current. At Eitel-McCullough, this relation, combined with the power-output versus modulating-anode-voltage characteristics of a klystron has been used to permit the generation of gaussian pulses.

Figure 6 shows both the block and functional diagrams of a shaped-pulse amplifier and transmitter system which has been developed to provide r-f excitation to the cavities used in the Cambridge Electron Accelerator's 6-Bev synchrotron. This system is an amplifier rather than a pulse generator, since it is capable of linearly reproducing any monotonically rising pulse generated at ground level by the input function generator. The waveform shown in the lower left-hand corner of Figure 6 is representative of the typical waveform required from this pulse amplifier. General specifications for the amplifier and transmitter are shown in Figure 6.

The block diagram of this transmitter system is shown in Figure 6A. The similarity between this block dia-

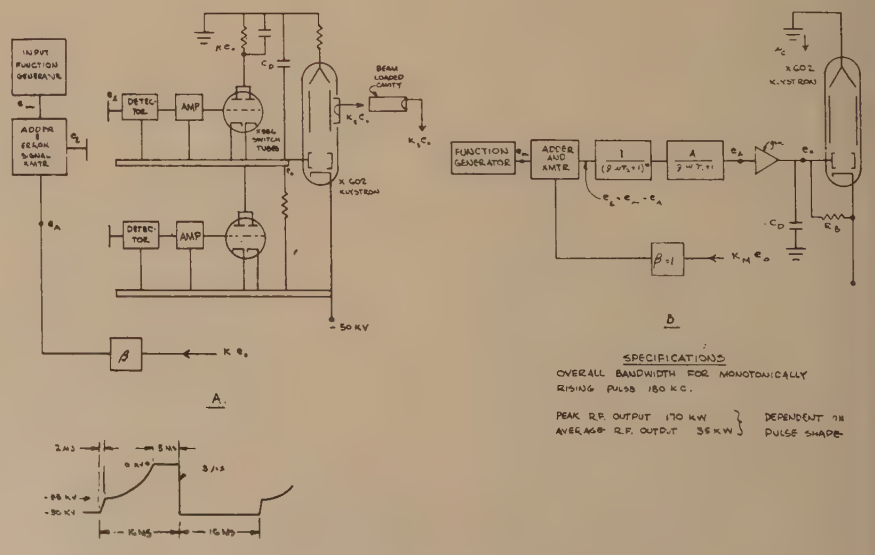


Figure 6 — Shaped pulse amplifier and transmitter.

ram and that of the basic floating-eck circuit is evident. In Figure 6A, ie switch-tube and tail-clipper-tube rive circuits have been replaced by n r-f detector and linear amplifier. 'hese amplifier circuits are so phased iat positive modulation of the r-f irrier transmitted from the error-gnal transmitter will turn on the -556 switch tubes, and negative iodulation of the r-f carrier turns on ie X-556 tail-clipper tubes.

With this provision, the entire mod-:ator could be operated as a closed-op amplifier. In the final version of iis circuit, negative feedback was de-ved by sampling the modulating-iode voltage by means of a capacitive ivider. During its initial design ages, this circuit was considered in rms of linear feedback theory and elded the functional diagram shown Figure 6B. In this figure, the series combination of the switch tubes and tail-clipper tubes is considered to be a single current source capable of passing current in either direction and the output voltage e_o is equal to the integral of this switch-tube current divided by C_d.

It must be recognized that the linearity assumptions involved in this discussion are violated in at least three different ways: First, since the X-556 switch tubes are normally at or very close to their cut-off voltage, the transconductance of these tubes varies in accordance with the error signal present. Second, since these high-voltage tubes are not ideal pentodes, their cut-off voltage varies as a function of the plate-to-cathode voltage across the tube. Finally, charging current effects which are unavoidably present at the driver-amplifier low-level stages, have to be phased in such a way that they are degenerative, and therefore the loop gain of the system varies slightly as a function of de_o/dt.

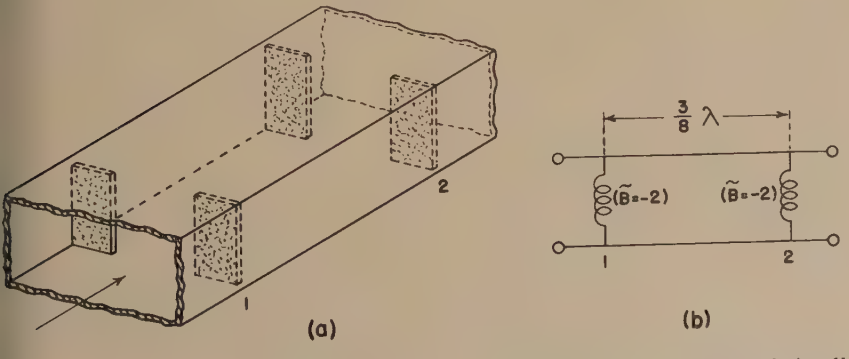
All these difficulties would have to be carefully investigated if one wanted to make a closed-loop amplifier which would reproduce other than monotonically rising pulses. For this case, however, these difficulties may be bypassed by the expedient of biasing the tail-clipper tube so that it only responds to large negative error signals. This permits one to raise the system loop gain without fear of oscillation, and use the resulting leading-edge overshoot to compensate for the effect of changing transconductance. In practice, the 10-90% step function response of this system was 2 μsec and pulse overshoot varied from 2% to 6% as pulse voltage was varied from 0 to 50 kv.

GRAPHICAL ANALYSIS OF MICROWAVE COMPONENTS

'ART II

George C. Southworth

his is the second of two articles on the analysis of microwave components by the Smith Diagram. In the last article, :amples were chosen centering around waveguide junctions. Numerous special cases were considered. One of special :terest incorporated symmetry. In it were the principles of the familiar device known as the waveguide stub-tuner. he present article will be concerned mainly with filters, but it will include also a device that, in effect, enables us to :ach inside an evacuated electronic device to perform experiments and thus learn much about its rather complicated :havior.



igure 1 (a) — A particular form of resonant chamber filter; (b) transmission-line equivalent.

We are indebted to D. Van Nostrand Co., Inc. and to Bell Telephone Laboratories for use of material in this paper.

Waveguide Frequency Filters

One of the more interesting as well as useful applications of graphical analysis lies in the field of microwave filter design. This comes during the earlier and exploratory stages of design. Some simple examples follow. Perhaps the simplest of all filter elements is the single resonant cavity, such as illustrated by Figure 1. Most of the facts relating to this simple arrangement may be visualized rather easily by means of an admittance chart presentation of the kind shown in Figure 2. It is assumed as a numerical example that inductive irises are used, each having susceptances of $\tilde{B}_1 = B_2$

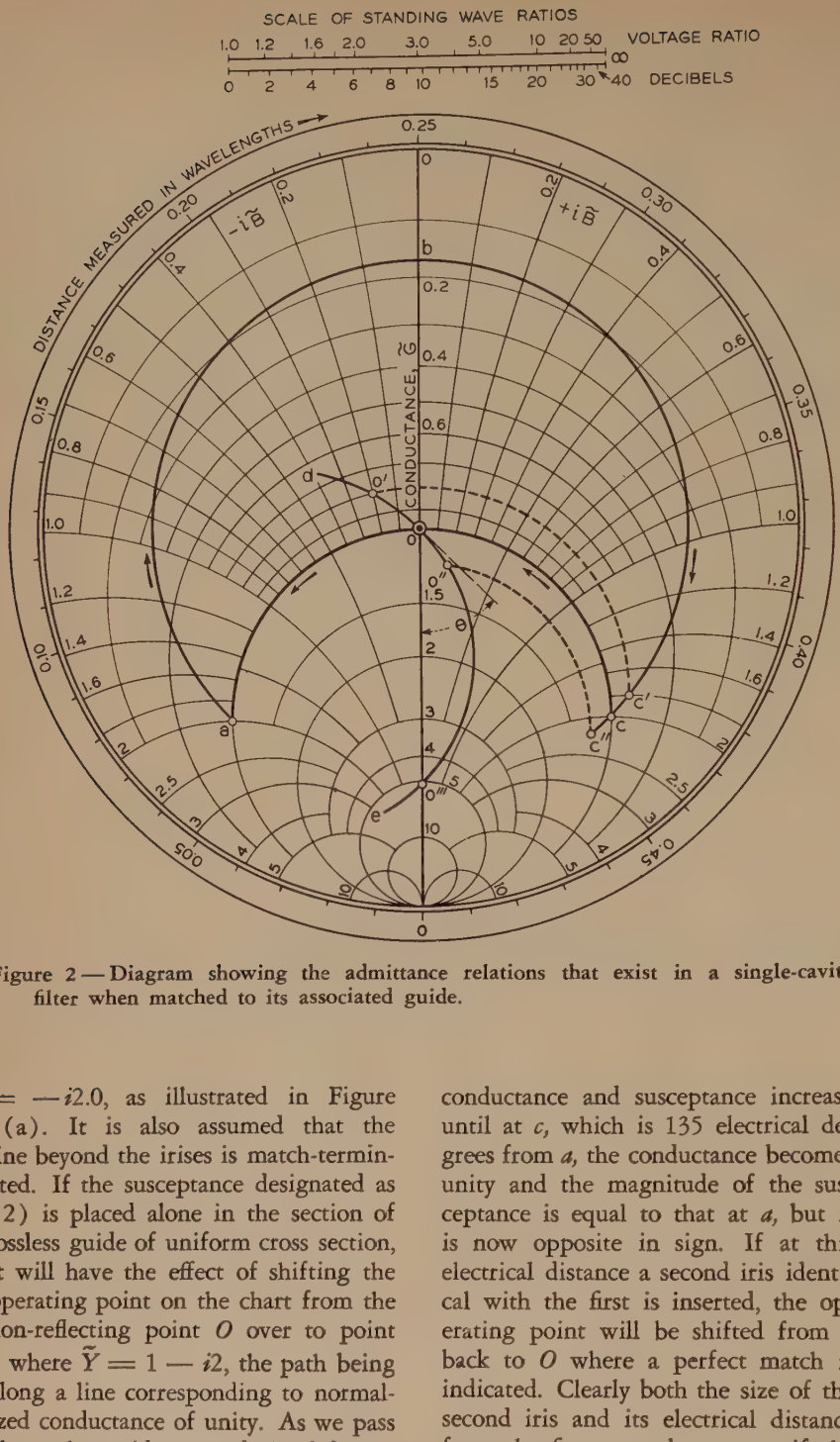


Figure 2 — Diagram showing the admittance relations that exist in a single-cavity filter when matched to its associated guide.

$= -i2.0$, as illustrated in Figure 1(a). It is also assumed that the line beyond the irises is match-terminated. If the susceptance designated as (2) is placed alone in the section of lossless guide of uniform cross section, it will have the effect of shifting the operating point on the chart from the non-reflecting point O over to point a where $\tilde{Y} = 1 - i2$, the path being along a line corresponding to normalized conductance of unity. As we pass along the guide toward the left, the apparent admittance will change as specified by points on a circle of fixed standing wave ratio designated on the chart as the path *abc*. Between the points *a* and *b* both the conductance and susceptance become smaller until at *b*, which is $3\lambda/16$ (67.5 electrical degrees) away from the first iris, the susceptance will be zero and the conductance will be a minimum. At this point then, the admittance of the iris, together with the match-terminated line, appears as a pure conductance, less than characteristic admittance. Continuing on toward *c* both the conductance and susceptance increase until at *c*, which is 135 electrical degrees from *a*, the conductance becomes unity and the magnitude of the susceptance is equal to that at *a*, but it is now opposite in sign. If at this electrical distance a second iris identical with the first is inserted, the operating point will be shifted from *c* back to *O* where a perfect match is indicated. Clearly both the size of the second iris and its electrical distance from the first must be correct if the operating point is to return to *O*.[1]

Having arrived back at point *O* the effect of varying the operating frequency will be that of sliding point *O* along a curve *de*.[2] This curve, which is the arc of a circle, is particularly significant, for it provides important information about the frequency characteristics of the chamber. To illustrate, assume that the operating frequency is reduced by a small increment such that the electrical length of the chamber will be decreased so that *c* moves to *c'* and *O* moves to *O'*. Thus the operating point moves into a region of inductive susceptan
This corresponds to the introduct
of insertion loss. Increasing the
quency has the effect of increasing
electrical length of the cham
thereby moving *c* toward *c''* and
toward *O''* and hence into a regio
capacitive susceptances. Again in
tion loss is introduced. From th
data we may predict the shape of
response curve. It will be noted t
in an extreme case, an operating p
O''' may be reached where the o
all load will again appear as a p
conductance, but this time it is
very substantial magnitude. This
ond point is also of practical imp
tance and will be considered ag
later. Returning to the matter of
quency response, if we wish to
crease its sharpness we should l
for means of increasing the angl
at which the curve *de* crosses the c
ductance axis. This may be accc
plished by choosing larger normali
susceptances (smaller irises), ther
transferring the circle of const
standing wave ratio to the ou
reaches of the chart. This corresp
to an increase in the length of
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The variation of admittance w
frequency of the single cavity str
ture may suggest to the reader ot
uses for such a device. That such u
are possible is evident from Figure
where it is assumed as before that
inductive iris having a normalized s
ceptance of $\tilde{B} = -i2.0$ is introduc
into a waveguide at an arbitrary lo
tion along its length. This again
the effect of transferring the operati
point from *O* to a point *a* where
admittance $\tilde{Y} = 1 - i2$. At a s
cified distance from this iris, cor
sponding to a point *c'* on the const
standing-wave ratio circle *abc'*,
apparent admittance of this iris
be approximately $\tilde{Y} = 0.8 + i$
If, this time, we introduce an iris h
ing a susceptance of $\tilde{B} = -i1.8$ (
$\tilde{B} = -i2$ as before), the operat
point will be transferred to a p
O' where $\tilde{Y} = 0.8 + i0$. At
point the structure should be able
match a pure normalized conducta
of 0.8.

If, on the other hand, we cho
an electrical distance from the first
corresponding to a point *c''* where
$= 1.2 + i2.2$ and if, at this point,

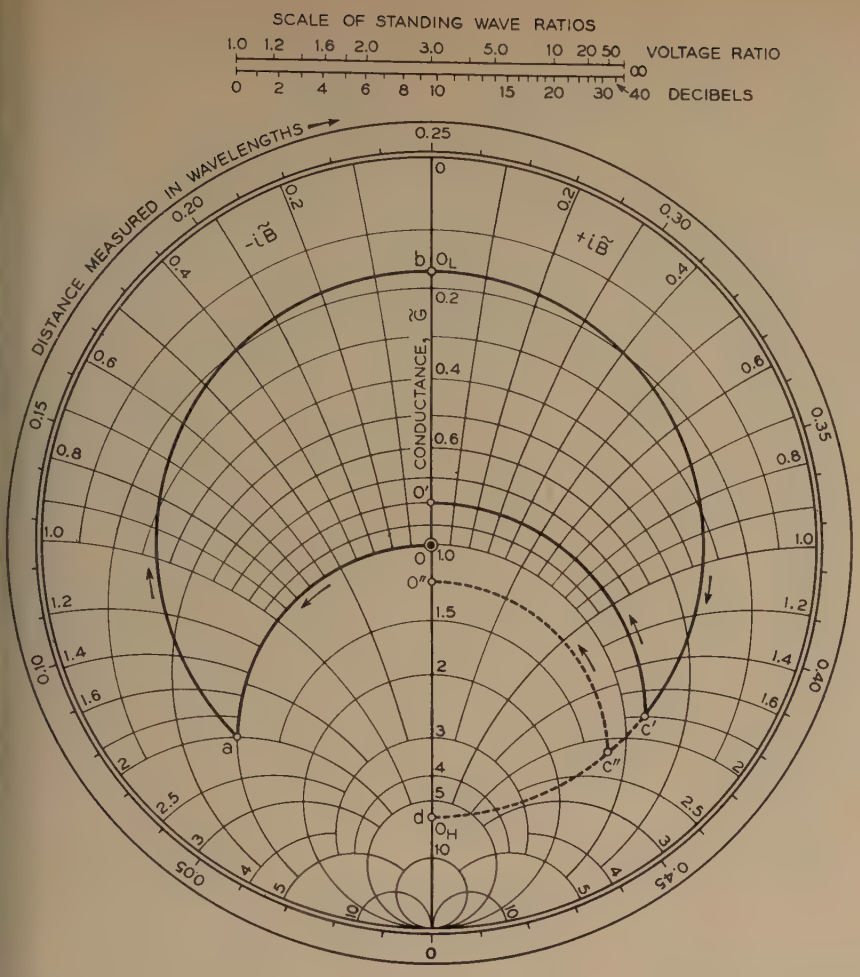


Figure 3 — Diagram showing how a single filter element may be used as a matching unit.

ris is introduced having a susceptance of $\tilde{B} = -i2.2$, then the operating point will have been shifted to a point $O'' = 1.2 + i0$. At this point the device could match an external conductance of 1.2. Thus the single cavity structure may, in addition to serving as a filter, serve also to match a waveguide of a given characteristic conductance to a conductance of some other magnitude. The latter may be another waveguide of different proportions or it may be another resonant chamber similar to the one now under consideration.

Further reference to Figure 3 shows that the conductances can never be less than or greater than those specified by points O_L and O_H, respectively. However, at these limiting points no supplementary susceptances (irises) will be needed. It also follows that, by replacing the first susceptance by another of greater magnitude, the circle of constant standing wave ratio will be transferred to the outer reaches of the diagram and accordingly the range between O_L and O_H will be increased. In the limiting case, the susceptance of the first iris may be made infinite (iris now closed), whereupon the path of constant standing wave ratio will be transferred to the periphery of the diagram. If, in addition, we imagine the conductance to be matched, to have been transferred to the interior of the chamber, the filter will then become a resonant-chamber type of transformer.

The Two-Chamber Filter

It has already been mentioned that a number of resonant chambers may be spaced along a uniform transmission line to provide a filter arrangement having a multiplicity of resonances. There is, in general, one fundamental resonance for each chamber provided. For the two-chamber case, which is to be considered next, there will therefore be two of these fundamental resonances. Arrangements of this kind are of considerable practical interest. Like two resonant circuits operated at critical coupling, they may present a low insertion loss extending over a fairly wide band of frequencies.

We shall approach the two-chamber case by considering initially two identical single-cavity filters of the kind just described, each perfectly matched in both directions to its adjacent waveguide line and each independently tuned to the same frequency. If the two filters are placed in tandem but are connected by a line having considerable attenuation, their respective transmission characteristics will still be identical. However, upon bringing the two filters into proximity, a considerable amount of interaction will take place; there will be component reflections from a total of four discontinuities. These four discontinuities are represented schematically in Figure 4(a) by the four susceptances B_1, B_2, B_3, and B_4. If finally we bring the two chambers into very close proximity, then the reflections from the two closely adjacent irises will tend to fuse into a single reflection. The resulting arrangement is represented schematically by the susceptances in Figure 4(b). The resultant susceptance is now the sum of the component susceptances at B_2 and B_3. In the waveguide case, this result may be accomplished by replacing the two component irises by a single iris (2) having a considerably smaller opening as illustrated in Figure 4(c). If there are no substantial losses within the filter, the susceptances of the two irises (1) and (3) will be equal and the susceptance of (2) will satisfy the relation

$$\tilde{B}_2 \doteq \tilde{B}_1^2 \doteq \tilde{B}_3^2 \qquad (1)$$

Iris (2) may be regarded as a coupling between the two adjacent chambers and, accordingly, we may refer to this arrangement as an iris-coupled filter. It is possible to visualize the phenomena involved in the two-chamber filter from the admittance diagram shown in Figure 5.

Assume as before that the end iris (1) of Figure 4 has the rather arbitrary normalized susceptance of $\tilde{B}_1 = -i2$. This susceptance alone has the effect of shifting the operating point on the diagram from O to a. Proceeding to the left from iris (1), one traverses, on the admittance diagram of Figure 5, a path abc. If at point c, we assume that an iris (2) is introduced having a susceptance $i\tilde{B}_2$ considerably larger (opening smaller) than the first, a variety of effects may occur depending on the magnitude of $\tilde{B}_2$ as well as its electrical distance back to $\tilde{B}_1$. Certain possibilities are

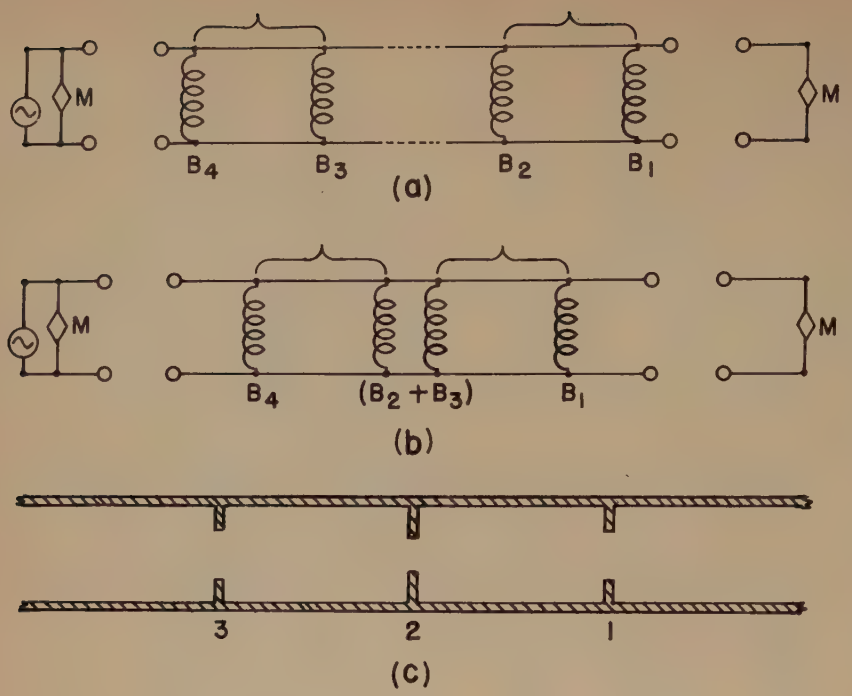


Figure 4 — A typical two-chamber filter with iris coupling. (a) and (b) transmission line equivalents for two representative degrees of coupling; (c) sectional view.

represented by the loci *gh*, *ij*, and *kl*. If, for example, one were to introduce a susceptance $\tilde{B}_2 = -2.8$ at a point *c*, it would have the effect of shifting the operating point from *c* to *d*, a point corresponding to a pure conductance (normalized) of about 3.0. At this point, then, the two-iris combination ($\tilde{B}_1 = i2$ and $\tilde{B}_2 = -i2.8$) is appropriate to match another pure conductance of magnitude $\tilde{G} = 3$. As we have already noted, one way to provide such a conductance is to add to the first chamber a second chamber of identical proportions. We shall then have a two-chamber filter of the kind under discussion. Continuing the admittance-diagram analysis through the second chamber, the path leads along the dotted line from *d* to *e* and thence clockwise to *f* where the susceptance $\tilde{B}_3 = \tilde{B}_1$ returns the operating point to *O*, which corresponds to the condition of perfect match with the line beyond.

It will be noted that the second part of the diagram is the mirror counterpart of the first and, accordingly, any combination of susceptances and electrical lengths which terminates the first part of the diagram on the conductance axis will, when followed by an identical structure, lead to a perfect match. Combinations which lead to a small residual susceptance in the vicinity of *d* cannot lead to a perfect match, but there may be cases, as we shall soon see, where the mismatch may

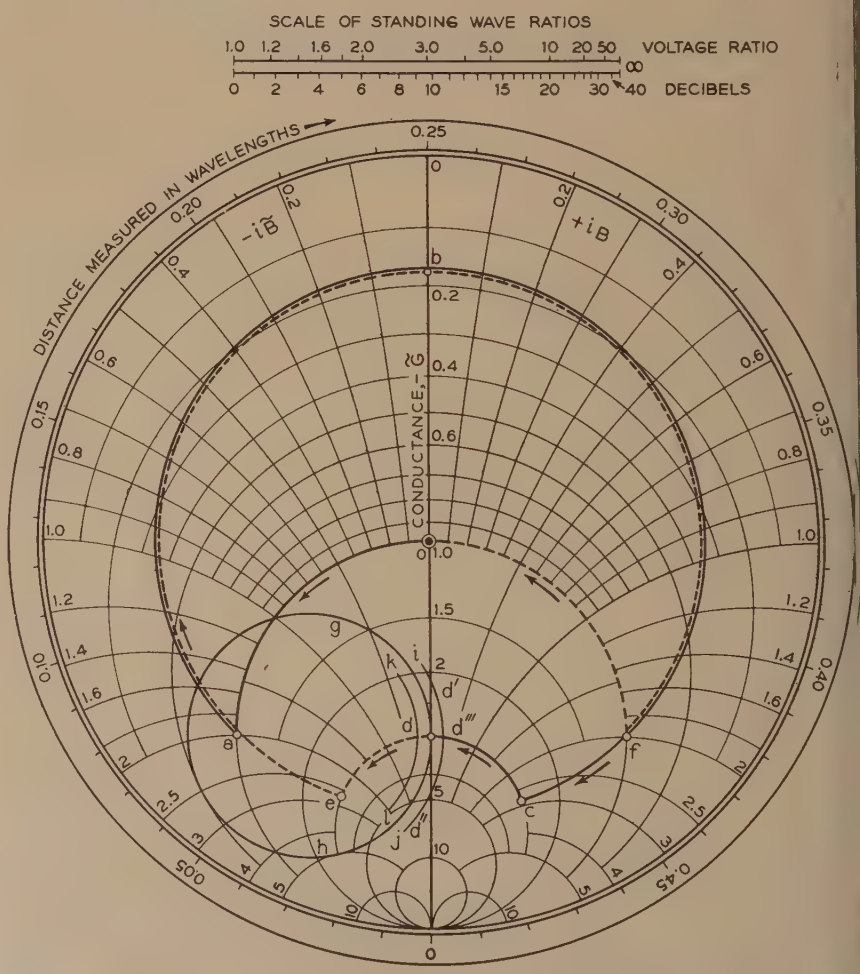
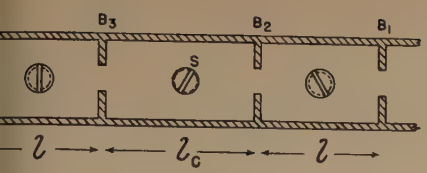


Figure 5 — Admittance diagram illustrating the properties of a two-chamber filter f three typical conditions of coupling.

Referring again to Figure 5 examining further the effects of e trical distance and the magnitude $\tilde{B}_2$, let us assume that they are b such as to terminate the first part the diagram at point *d* on the cond tance axis. If, in this case, we imag the frequency to be increased b small amount, thereby increasing electrical length of the first chamb point *d* will move clockwise arou the locus of points *gh*. Decreasing frequency naturally leads to an op site effect. If the frequency shift small, a correspondingly small susc tance will be incurred either side point *d* but, at other more rem frequencies, the operating point m move further around the circle *gh* a the admittances may become v large indeed. Because, in the vicin of *d*, only small susceptances are curred by moderate increments of c frequency operation, we may conclu that the frequency characteristic w



gure 6 — A filter consisting of two identical chambers coupled by a third chamber.

broad. Also the frequency characristic will have a single peak.

Next let us assume that the magnide $\tilde{B}_2$ is reduced by a small amount as to lead to a locus of points repreted by the circle *ij*. The conducnce axis is now cut at two points. etween these two points, a small but vertheless appreciable susceptance is curred, whereas for frequencies outde these points the magnitudes of the sceptance component is at first ther small, but it soon becomes conderable and increases rapidly as we oceed toward more remote points on e locus. We conclude from this relt that the frequency characteristic rresponding to this case will have o peaks with a small depression beeen. In the region outside the two aks the characteristic will have ther steep sides. It is obvious too at, by a suitable manipulation of e magnitude of the coupling susceptnce $\tilde{B}_2$, we may vary the distance tween the two peaks at will, effectg thereby a large variety of comomises between bandwidth on the e hand and midband insertion loss the other.

Finally let us assume that the magtude of the susceptance $\tilde{B}_2$ is now eater than 2.8. The locus of points r *d* will be a third circle *kl*, which ay approach the conductance axis t never quite reach it. Since the sceptance may be a minimum but ver quite zero, we conclude that ere can never be a perfect match d hence there will be considerable sertion loss even at the middle of e band. The frequency characteris- will have a single peak.

vity-Coupled Filters

Certain filter combinations may asme, under special circumstances, aracteristics ordinarily found in a ther different kind of structure. For ample, it is possible to construct a ree-cavity filter of such proportions at one of the three peaks, ordinarily pected, is pushed outside the observle range of frequencies. In this case, ly two resonances are evident and accordingly the arrangement assumes properties ordinarily found in a two-cavity filter. As we shall later see, the middle cavity serves as a coupler between the cavities on either side and, for that reason, this arrangement is known as a cavity-coupled filter. Although such a structure may appear, at first sight, somewhat more complicated and may actually occupy considerably greater space than a two-chamber filter of the kind just described, it has certain very desirable features.

As a convenient approach to this problem, it is well to return to Figure 4 and assume as before that two resonant chambers, adjusted to the same frequency, are placed in tandem. However, instead of merging the two chambers into a single two-cavity structure, we shall separate the two cavities by a distance slightly less than a quarter of a wavelength.[3]

The resulting structure, therefore, will consist of three chambers such as shown in Figure 6. Under this circumstance the component reflections from the four irises will develop a second resonant frequency adjacent to that for which the two chambers were individually adjusted.[4]

The conditions prevailing in a cavity-coupled filter may be visualized by referring to the admittance diagram shown in Figure 7. The locus C_1 summarizes the characteristics of the chamber between B_1 and B_2 when viewed from a point immediately to the left of B_2. The arrow indicates the direction of change of admittance as frequency is increased. It will be noted that the locus C_1 crosses the conductance axis at point O which corresponds to a match.

If an observer located in the guide moves away from iris B_2 a distance of one-eighth wavelength, the susceptances that he will see looking back toward the iris B_2 will be as if the arc C_1 had been rotated into the position indicated by arc C_2. This will cut the conductance axis not only at the original point O but also at another point P, and the interval between O and P may be varied merely by rotating C_2. Thus P may be located either at O or above or below O. The condition for which P falls on O is

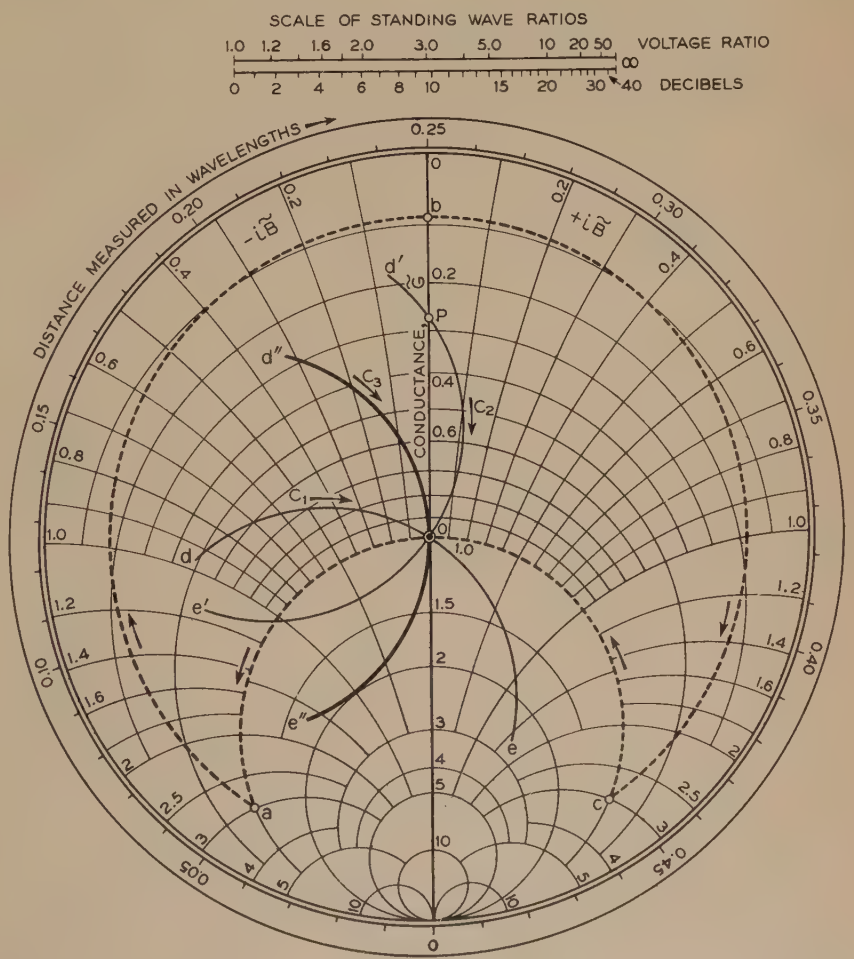


Figure 7 — Diagram showing the admittance characteristics of the cavity-coupled filter shown in Figure 6.

represented by the circle C_3. This corresponds to a distance somewhat less than one-eighth wavelength.

If, at the point along the waveguide structure at which this condition prevails, a second identical structure including another similar spacing is added, then the admittance as seen in both directions will be the same, and an over-all match will result. It is particularly significant that, whereas perfect match prevails at one frequency only, the mismatch prevailing at neighboring frequencies may be acceptable even over a fairly wide range. This corresponds to the maximal condition already referred to.

Although the maximal condition provides a fairly wide band of frequencies, it is possible, by accepting certain compromises, to extend this band very materially. For example, the length l_c of the coupling chamber may be made either slightly greater or slightly less than that corresponding to the maximal adjustment, thereby providing two near-by points O and P at which the locus C_3 crosses the conductance axis. Since both points correspond to a perfect match and since there is appreciable region between the two points as well as the regions outside over which the mismatch may be acceptable, the available bandwidth has been materially extended.

The Rieke Diagram

It often happens that when a useful invention is made, applications are found that were not anticipated by the inventor. We suspect that this was the case in the very useful application often referred to as the Rieke diagram.[5] The latter provides very informative data about the operation of microwave oscillators, as for example, magnetrons. In collecting data for a Rieke diagram it is customary to make observations at some arbitrary point along the output line, preferably not too far from the oscillating source itself, with apparatus such that frequency and power can be measured at each of a fairly large number of loads. Such loads must vary not only in conductance G but in susceptance B as well, and in particular, provision must be made for measuring the magnitude and position of the standing wave relative to the load. These latter measurements provide information about the coefficient of reflection as well as its phase and hence the admittance (normalized) into which the oscillator operates. Sometimes the susceptive component is produced by a device known as a standing wave producer. This device is much like a standing wave detector except that the probe is introduced into the slotted guide to the point where the required standing wave is produced.

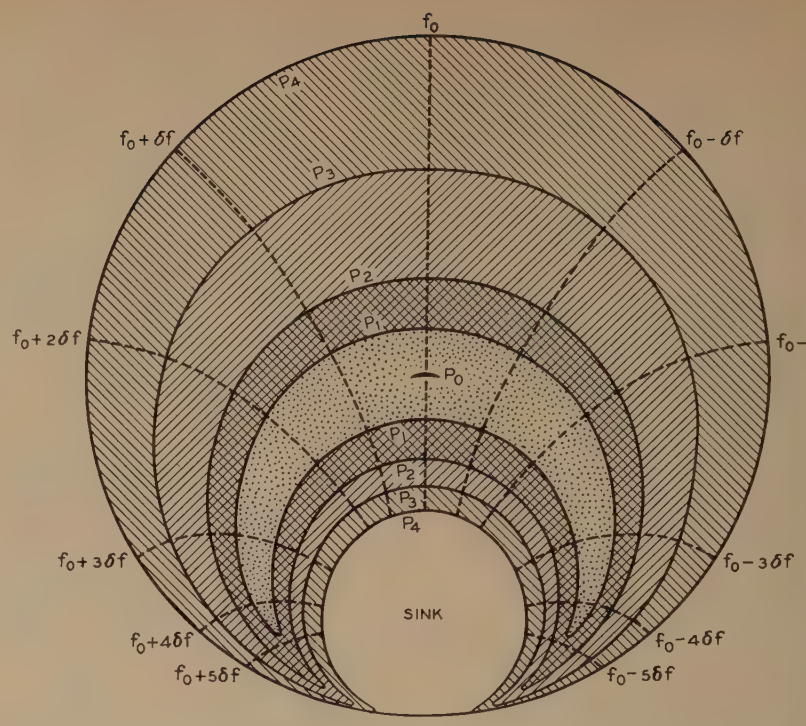


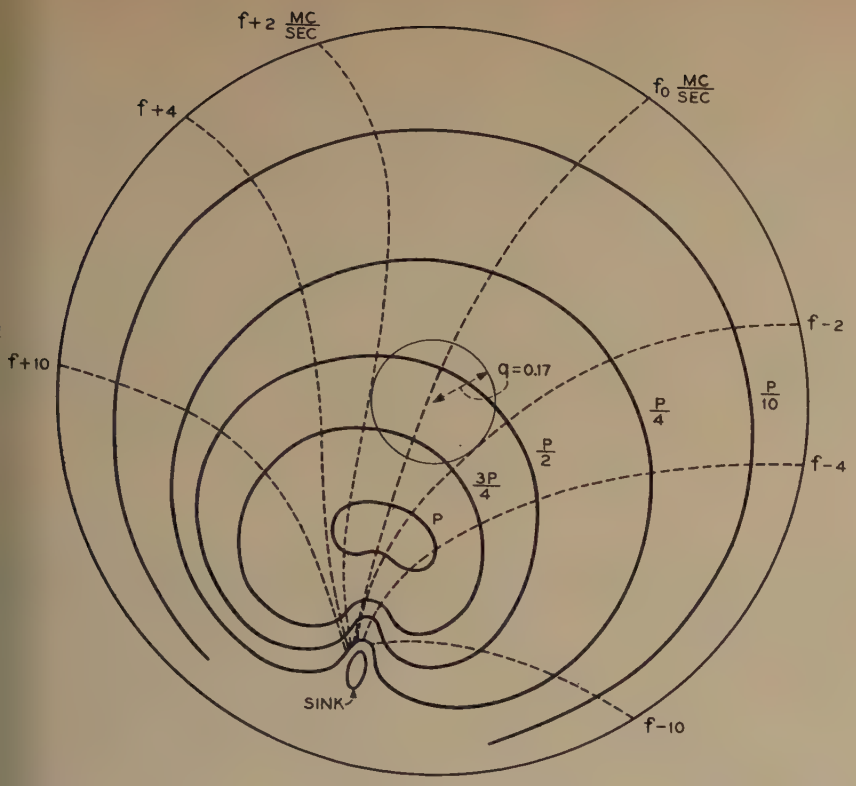
Figure 8 — Idealized form of Rieke diagram showing lines of constant power and lines of constant frequency. It is assumed that $P_1 = 0.94\ P_o$, $P_2 = 0.78P_o$, $P_3 = 0.44P_o$, and $P_4 = 0$.

Using for the moment only the angle and standing-wave ratio grid of the Smith Chart, magnitudes of standing wave ratio are plotted against electrical distance (angle) from the nearest minimum to the load, noting on the chart, at the same time, the power and frequency. After a sufficient number of points have been plotted, it will be possible to draw on the diagram contours of constant power and also contours of constant frequency.

There is shown in Figure 8 a much idealized Rieke diagram applying to a reflex oscillator. It is assumed that coupling to the load has been carefully adjusted so that optimum output corresponds to a match with the external load (center of diagram). In order to avoid confusion, the background grids of the Smith diagram have been omitted. The latter are, of course, made up of lines of constant susceptance and constant conductance and also angle and standing wave ratio.

An important point on the Rieke diagram is that which represents optimum power P_0. For the particu conditions here assumed, this occ at the center of Figure 8. Outside point there may be drawn other c tours shown as solid lines, each c responding to lower levels of pow Four such contours are presented, c responding respectively to P_1 $0.94P_0$, $P_2 = 0.78P_0$, $P_3 = 0.44$ and $P_4 = 0$. Each contour inter corresponds approximately to two cles of normalized conductance the background grid. A contour special significance is $P_4 = 0$ si it represents the locus of points which oscillations cease. Reference the admittance grid beneath sho that, for the particular conditions h assumed, this corresponds to the t circles $\tilde{G} = 0$ represented by the pe phery of the diagram and $\tilde{G} \doteq 2$ r resented by the small circle near bottom of the diagram. The area side this circle represents values conductance (and susceptance) which no oscillations are possible.

For each point plotted on the d gram, there is a corresponding f quency at which the assigned pow will prevail and for which the electr admittance plus the circuit admittan (including load) are equal to ze When these corresponding values frequency are plotted on the Rie diagram, we may then draw in lin of constant frequency. These are re



igure 9 — Typical Rieke diagram plotted from experimental data. The data apply specifically to a reflex oscillator.

esented by the dashed lines of Figure . They are approximately orthogonal the lines of constant power. It will e noted that they converge in the egion near the bottom of the diagram here no oscillations are possible. This egion is sometimes known as the sink."

The frequency for which optimum utput power is produced is of conderable interest. In the more-or-less leal condition here assumed, the line epresenting this frequency corre-onds to the zero susceptance axis n the background grid. It is, there-re, the line corresponding to zero usceptance in the load. In practice, is found by varying the susceptance the load until the output is a maxi-um. When so tuned, variations in e conductance term appearing in the ad will carry us along this constant frequency line to the point P_0 of optimum power output.

Actual Rieke diagrams may depart markedly from the idealization shown in Figure 8. Seldom is the diagram symmetrical and seldom is the point of optimum output at the center of the figure. Indeed in practice the diagram may look quite unlike Figure 8. Yet most of its essential features may be identified. Figure 9 is typical.

An important figure of merit of an oscillator is its frequency stability. It is often specified as the ratio of departure of assigned frequency ∂f to the assigned frequency f_0. Thus $M = \frac{\partial f}{f_0}$. The Rieke diagram provides useful data for finding a satisfactory region for operation.

For many purposes it may not be desirable to operate in the region of maximum power output. To do so might carry the operating point into a region where the oscillator might be undesirably frequency-sensitive as regards load changes. Also there are good reasons for keeping well away from the sink. Having agreed upon a satisfactory figure of merit, a diagram such as Figure 8 points the way to the best operating point. Often this is remote from the point of maximum power output.

Summary

This is the last of several articles published in this Journal centering around the Smith diagram. Included were (1) some personal notes about the originator of the diagram, Mr. Phillip H. Smith; (2) the origin and evolution of his diagram; (3) an enumeration of the kinds of problems that may be solved by its use; (4) procedures to be followed in such solutions, and finally (5) illustrative examples of its use. The latter ranged in complexity. At one extreme, it was shown how complex impedances may be converted very simply into equivalent complex admittances. At the other extreme, it was shown how the Smith diagram may explain the functioning of microwave filters. Indeed it is the insight into the functioning of such devices that makes the Smith diagram so very useful. Intermediate between these two extremes were other examples, as for example, matching networks, illustrating how a maximum of received power transmitted along a waveguide may be impressed on a tiny receiver.

It is hoped that the examples cited will point the way for the student reader to many other problems, preferably of his own origin, leading not only to correct solutions but to a better understanding of how various devices operate. In engineering education, next to a knowledge of fundamentals, a sense of magnitudes is possibly the most important item.

FOOTNOTES

It is an interesting exercise to determine from the transmission chart the compromises that may be necessary if the irises are not lossless or if the attenuation along the section of waveguide cannot be neglected.

It may be shown that the curve de is the arc of a circle of radius $r = \frac{\tilde{B}\sqrt{1+\frac{\tilde{B}^2}{4}}}{2\left(1+\frac{\tilde{B}^2}{2}\right)}$ centered on a straight line passing through the origin and making an angle with the real axis of $\phi = \frac{\pi}{2} - \theta = \tan^{-1}\left(\frac{2}{-\tilde{B}}\right)$.

3. Because of fringing effects at irises and a corresponding interaction, a half wavelength is customarily added.

4. A very complete description of this filter is given by W. W. Mumford of the Bell Telephone Laboratories in the B.S.T.J., Vol. 27, pp. 684-713, October, 1948.

5. The Rieke diagram is said to have been devised during World War II by F. F. Rieke and J. E. Evans for testing oscillators and later was used rather generally for this purpose.

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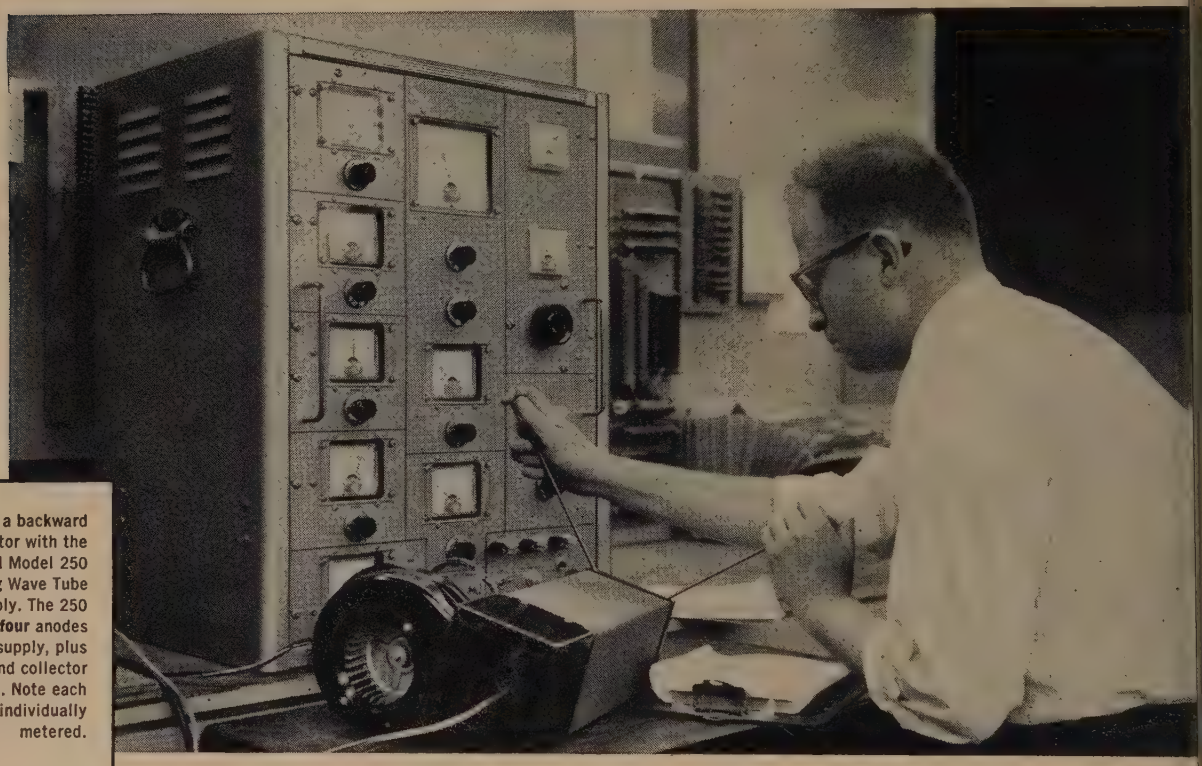

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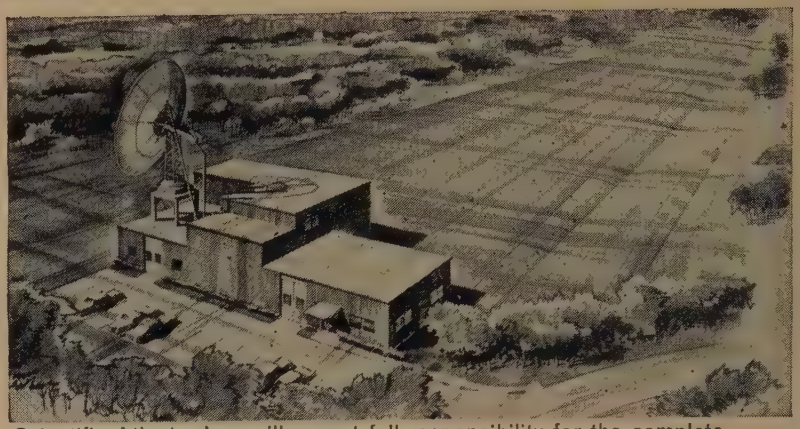

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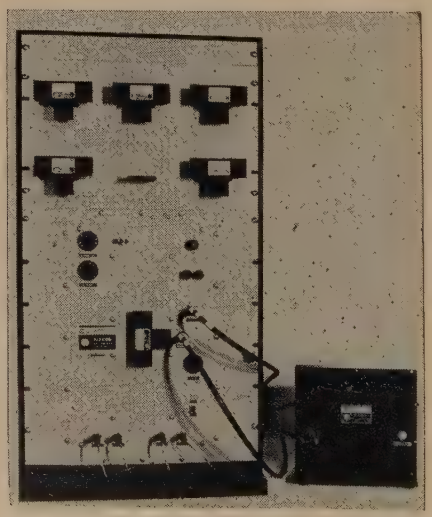

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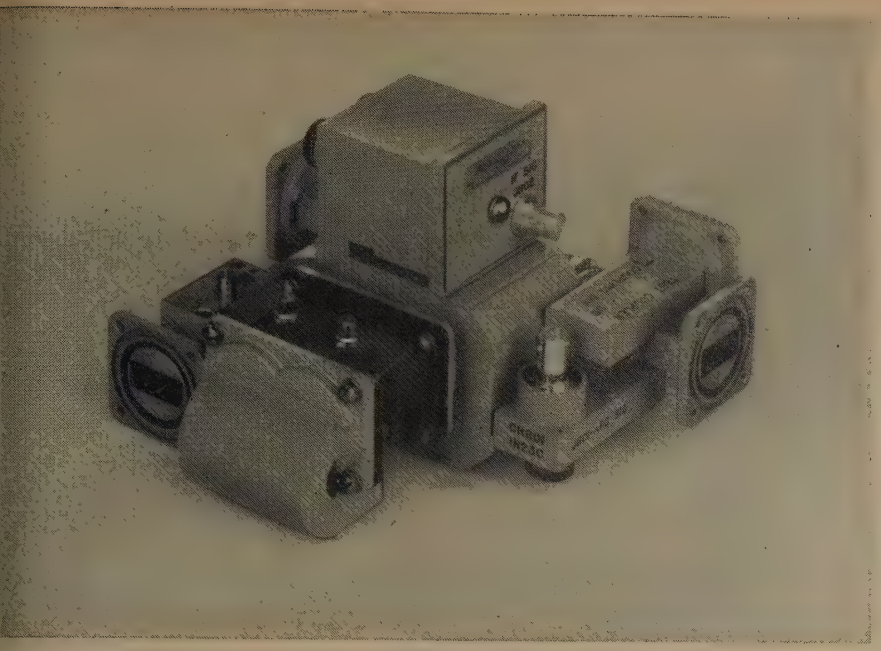

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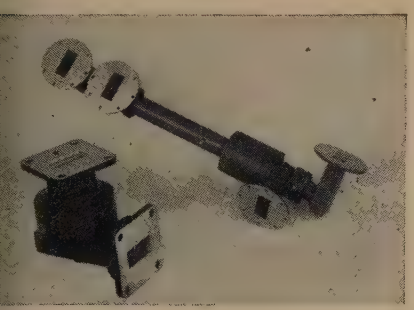

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		8.2-10.0	1.2						
90	X	8.4-9.6	1.1	1.02	.10	150 KW	In-Line	30 "	12 " "
		8.2-10.0	1.2						
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Insertion Loss	0.8 db max.
VSWR	1.15 both ends
Length	1.7"
Waveguide	RG 106/U
Weight	9 oz. (UG-441/U)
	5 oz. (RETMA)

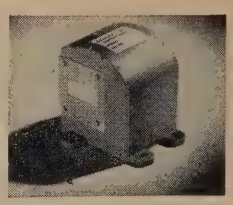
Coaxial Isolator

For TWT and Other Special Applications

Specifications for Model I-301L

Frequency	5000-6000 mc
Isolation	20 db min.
Insertion Loss	0.8 db max.
VSWR	1.20 max.
Connectors	Type "N"
Length	5¼"

Broadband High Power X_s ISOLATOR



For Rugged Military Specifications

Specifications for Model I-302H

Frequency	8200-11000 mc
Isolation	15 db min.
Insertion Loss	0.6 db max.
VSWR	1.15
Peak Power	250 kw
Average Power	300 watts
Length	2.5"

Ferrotec, Inc. announces new advances in the miniaturization of Ferrite Isolators! Send us your electrical and packaging requirements.

Ferrotec, Inc.
217 California Street
Newton, Massachusetts
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MELabs, Inc.
of Palo Alto, California

Artists view of MELabs, Stanford Park facility to be complet in late '59.

Introduction

Aside from being a young art, electronics is an art for the young. It is a mushrooming industry whose growth and direction is largely dictated by an expanding group of relatively young men. As a growing industry, intimately known by the young and daring, it exudes the small business spirit so integrally associated with the growth of our nation.

With technical capability, a reasonable approach to business problems, and an endowment of small business spirit, you, too, may be able to initiate a successful electronics organization. In doing so, it is necessary that you

1. Fulfill a need.
2. Have engineering competency.
3. Know the people and ways of the industry.
4. Be prepared to tackle new engineering and business problems.

Industrial giants are not invulnerable competitors in the rapidly advancing electronics field. Because of the giant's slowness to react to change, small businesses flourish and grow in their shadows, and prospective customers frequently find it easier to deal with a small business where there is a minimum of red tape. Frequently, however, it is not necessary to compete with large business, because many of them prefer to contract for special talent rather than build up a given technical specialty. Under the current weapons systems concept of military contracting, a large organization can be a powerful partner.

One of the strongest deterrents to the creation of sm business today is the individual's desire for security often associated with employment among the busine giants. The world abounds with prospective entrepreneu stifled by the cry for security. Security is not found nece sarily in the large company, because real security li within the individual. The cutbacks of late 1957 saw la offs at all employment levels and were most promine among some of the industrial giants. In a situation su as this, the most competent individuals are generally last be laid off although there are certainly exceptions. A capable individuals were able to find new employme during this period without undue delay. These peop possessed job security built up by job knowledge gain through self or formal education. Lacking fear of t unknown, these men have security whether working f the giant, for small business, or for themselves.

At MELabs, the philosophy is to create security with the man.* This approach is taken through assistance further education, by continually bringing him face to fa with the technological unknown, and by taxing his skil The apprehension of facing a frontier for the first ti is soon turned into a lust for technical conquest. Sm laboratories thrive on the R&D spirit of such men a MELabs is among them. Of the 130 staff members of t laboratory, over one-third are R&D engineers, and ma of the laboratories' 40 technicians are qualified for ind pendent development work.

*Editor's note: Amen.

History

MELabs grew from an idea and discussion generated board an eastbound transcontinental airlines flight in ebruary, 1956. Lloyd A. Addleman and Jack L. Melchor ere going to present papers at a New Jersey symposium. etween San Francisco and the Continental Divide, it be- me established through social conversation that each was nterested in the possibility of self-employment.

Lloyd, who headed the microwave receiver group at the lectronics Defense Laboratory had a broad background n electronics circuit development. A product of Ohio State Jniversity in 1949, he worked in radar and IFF equip- nent development at Hazeltine after graduation. As proj- ct engineer for the Press Wireless Manufacturing Co. uring the early 1950's, Lloyd assisted in the development f secure communications systems. In 1952, he joined the ylvania Physics Laboratory at Bayside, New York, working n circuit techniques for applications of cross-correlation o Loran, and was transferred to Mountain View a year ater.

Jack, who headed the microwave ferrite group at the lectronic Defense Laboratory, had a physics background. A graduate of the University of North Carolina, he had erved as Civilian physicist with the U.S. Navy Mine Countermeasures Station in Florida. In 1952-53, he worked n systems analyses at Bendix Aviation while completing work for his doctorate at Notre Dame. In 1953, he, too, oined the Electronic Defense Laboratory.

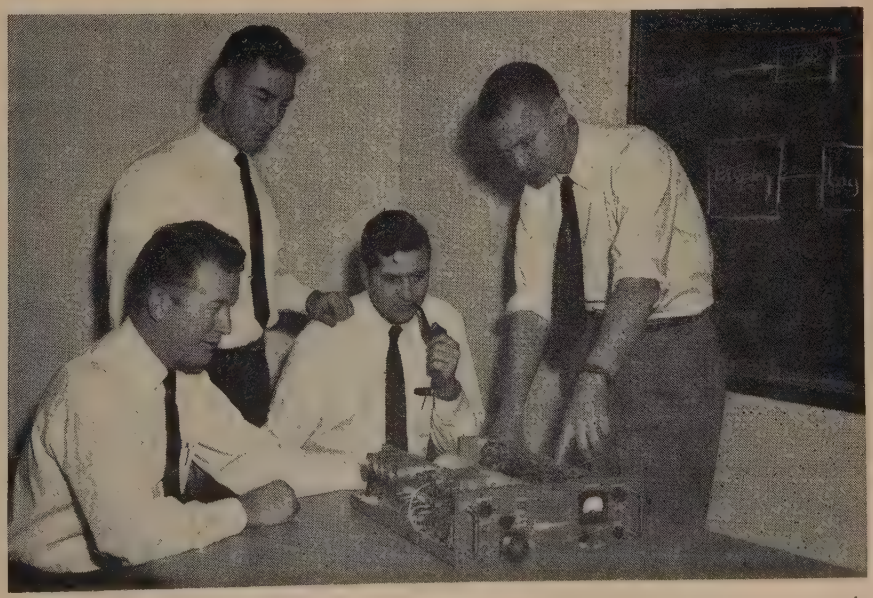
In the course of their work during the following three ears, both men had occasion to participate in develop- nents of mutual interest to the receiver and ferrite groups. Although they had not been close friends, the interaction ad provided an opportunity for personal appraisal, re- sulting in a feeling of mutual respect. Therefore, when he possibility of establishing an independent organization merged, each became enthusiastic about the prospects.

After landing in New York and driving to Belmar, New Jersey, conversation was dominated by the prospects of tarting a new business. Microwave electronics was not the nly field of interest considered. Discussion covered a wide range of possibilities, even including such things as lectronic toys. However, upon discussing government contracting with small business specialists, it was decided that an attempt would be made to break into military microwave electronics. A number of close friends attending the symposium were both helpful and encouraging when the subject of a new company was broached. The specialties involved here, intercept receivers and microwave ferrites, were achieving a degree of prominence in the military electronics field.

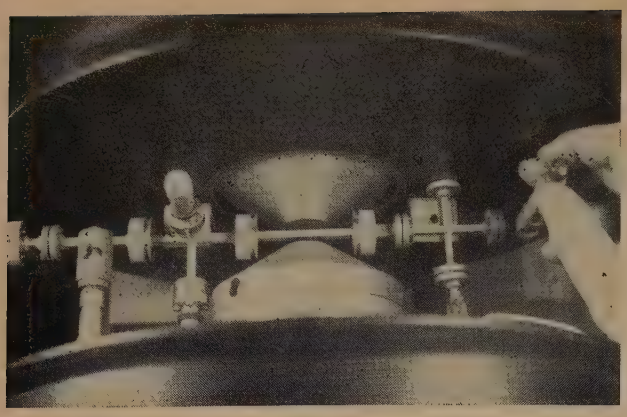
Upon conclusion of the symposium, Lloyd prepared to return to the West Coast while Jack remained in the New York-Washington area. At this point, the idea of a new venture was still in the stage of a verbal daydream, similar to that held by most young people during the course of a career. However, it was agreed that Lloyd would contact Perry Vartanian on his return to see if Perry might also be interested in such a venture. Perry was an exceptional young engineer, working in the ferrite group at EDL, who was just completing requirements for his doctorate at Stanford University. In his two years in microwave ferrites work he had exhibited outstanding ability and tremendous drive. As it turned out, Perry was immediately enthusiastic over the idea, and upon Jack's return to California, the three met evenings and week-ends to plan the initial stages of incorporating, procuring business, etc.

At this stage of development, it still was not certain that the dreams could be fulfilled. Although Perry was a bachelor, Lloyd and Jack each had a family with two children to be supported, and the gamble of resigning $10,000 a year jobs on the hope of future business was not an easy choice to make. A windfall developed about this time to ease the decision. Professor Bracewell of Stanford University was in need of an S-band radiometer for his radio astronomy program. Lloyd reviewed requirements with him early in April and prepared a proposal for submission in parallel with his resignation on April 15, 1956 to become the first member of MELabs. Professor Bracewell found the proposal suitable for his needs and a purchase order was issued May 15, the date of Jack's resignation from EDL to become the second member of MELabs. The first laboratory facility was organized in a two-room prohibition wine cellar on the Addleman property.

The second contract was for microwave ferrite switches



MELabs founders, left to right, Wesley P. Ayres, Perry H. Vartanian, Jack L. Melchor and Lloyd A. Addleman.



Microscopic examination of 2 mm waveguide components manufactured at MELabs.

Waveguide setup for ferrite 2 mm frequency doubler.

and drivers from Ewen-Knight Corporation. Although these two contracts guaranteed collectively less than $15,000, they provided the necessary support to start building an organization, and the company was moved to a 1,000 square foot area in half a plumbing shop. Meanwhile, overtures had been made to Wes Ayres to join the trio as a participant in the company. He was elected to the board of directors at its first official meeting following incorporation. Wes had been a member of the EDL ferrite group for two years following receipt of his doctorate at Stanford University. A fifth stockholder, Professor C. L. Hogan* of Harvard University, had also been invited to participate in the foundling company. Professor Hogan was known world-wide for his development and analysis of the microwave ferrite gyrator. Collectively, these five put up $22,600 for initial capitalization.

Government contracts, which were expected in the Summer and Fall of '56 never materialized and it became apparent that additional sub-contract support would be necessary. Litton Industries, having heard of the organization, sent representatives in for contracting and acquisition discussions during midsummer. Since MELabs was in its infancy and any acquisition would defeat the total purpose of going into business, this was not considered seriously. However, negotiations were conducted for several months regarding MELabs microwave ferrite consulting and contracting. These were discontinued in the Fall, but by then Airtron, Inc. had become interested in an exclusive consulting and development contract with MELabs. Preliminary discussions began with Airtron at the Los Angeles Wescon meeting in August, and a final two-year contract was signed in Cambridge, Massachusetts October 8, 1956. This was an extremely timely contract in the growth of MELabs. Both Lloyd and Jack had completed consulting contracts with Sylvania, and MELabs had become their sole support. In addition, Don Gardner, who was the first bonafide employee had joined the company in August and

*Now manager of the Motorola Semiconductor Products Division in Phoenix, Ariz.

Wes Ayres, the fourth founder, had joined it October
The Airtron contract guaranteed development wo
which stabilized the MELabs engineering effort whi
providing much needed development capability to A
tron. As has since been voiced by both parties, t
Airtron-MELabs contract was one of great mutual benef

MELabs prime government contracting began in Apr
1957 with a six months study of the solid state maser. Pr
fessor Hogan was instrumental in making known the la
oratory's capability in solid state microwave work to A
Force personnel in the interim between Bloemberger
maser invention and his first paper on it in the Physic
Review. Contract negotiations which began in the F
of 1956 culminated in the April, 1957 contract. Dela
had been encountered while MELabs established the nece
sary financial and technical capability that are prerequisit
to government contracting. Shortly after the first contra
award, several additional equipment and solid state co
tracts were received from government agencies, and th
company was on its way to becoming established in m
itary electronics.

MELabs' Technical Activities

Today, people frequently express surprise at MELab
variety and extent of technical activities. Since its ince
tion, MELabs has developed a number of radiometer
and ferrite devices ranging from L-band to 2 mm, inclu
ing the 1400 mc ferrite choppers used in many of th
world's major radio telescope installations. Circuits d
velopments have included a variety of microwave receive
including the first successful imageless superhet microwa
receiver, special storage and display equipment for th
military, various signal sources, and diode parametric u
converters. Component developments include a varie
of octave bandwidth stripline, coaxial and waveguide co
ponents, including broadband variable directional couple
mixers, improvements on many standard components, a
a multitude of active and passive ferrite devices. So

tate microwave studies include broadband masers, ferro-nagnetic parametric amplifiers, millimeter wave genera-ion in ferrites, (the 50 watts peak output so far ob-ained is the highest reported at 2 mm), uses of para-nagnetic resonance in microwave receivers, ferrite single ideband modulators, and general phase shift studies of errite loaded structures.

With this range of technical activity in process, MELabs s beginning to embark on systems programs of military mportance. In the future growth of the company, it is nticipated that military systems will play a vital role.

MELabs Growth Record

Starting with $22,600 in capital, MELabs has grown o a net worth of almost $200,000 in less than three years; ased on recent trends, the public stock market would alue the company in excess of $1,000,000. In this three-ear period, sales have increased from the initial radiom-eter contract for less than $10,000 to a rate in excess of $1,500,000/year. Although the laboratory holds a half dozen prime contracts, its principal sales come from ubcontracting to some of the larger companies such as Stromberg-Carlson of General Dynamics and the Sperry Gyroscope Company.

As the company's first employee, Lloyd Addleman has een it grow steadily to its current 130 persons. Over two-hirds of these are technical personnel, but the percentage s decreasing with the weekly addition of fabrication and ssembly personnel. Approximately, two-thirds of the ompany employees are stockholders while others are waiting the end of a 3-6 months stock eligibility period. Employees own in excess of 99% of the outstanding tock, and they are accumulating it at a fraction of its narket value. Participation in earnings and growth plus challenging work have built strong company loyalties among MELabs' personnel. The annual picnic or Christ-nas party finds them turning out en masse to continue trong social ties developed through the years.

A monthly laboratory meeting keeps all personnel in-formed of new business, new prospects, problem areas, deliveries, financial position, new employees, future plans, etc. A concerted effort is made to maintain the small busi-ness attitude while developing a larger business outlook. A monthly dinner-meeting of the senior engineering staff keeps them abreast of company-wide technical problems.

Having grown from the Addleman's wine cellar, via a plumbing shop, MELabs now occupies 18,000 square feet on Industrial Avenue in Palo Alto. Since the space is split between two buildings, the informality of seeing all of the co-workers daily is being lost among the employees. However, consolidation into a single building should be completed within the current year.

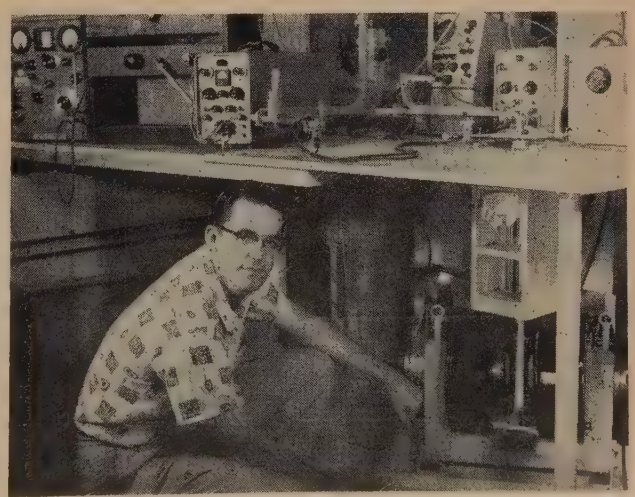
Future Forecast

A backlog in excess of $2,000,000 will dictate the con-inued expansion of MELabs. Extension of this backlog s fairly well insured by expanding national requirements in the laboratory's technical specialty fields. By some organizations and many people in the electronics field, military electronics is being overtly decried for sake of a "commercial" market. Such an attitude has developed due to the military's ability to stop buying on short notice. However, one should not be too hasty in discounting the military market, because discontinuance or reduction of military electronics purchases tends to dry up much of the "commercial" market, especially in electronics capital equipment. Companies that were expanding in military electronics suddenly stop acquiring test equipment until military business is resumed. There is little doubt that a prolonged stoppage of military purchases would create an overabundance of competition for the true commercial market with many companies fighting to take up production slack. Having seriously considered the consequences of military electronics, MELabs is willing, without hesitation, to tie its growth to the military needs. This does not mean, however, that commercial markets such as medical and industrial electronics will be ignored. It is anticipated that the commercial market will become a substantial part of MELabs future business, but military electronics will not be disparaged.

Plans are underway for constructing a new 32,400 square foot laboratory on the Stanford University campus. In the new facility, a greater percentage of space is being allocated to manufacturing. The site and building layout have been chosen to provide for orderly expansion in future years. There, it is anticipated that the MELabs story will develop until one day MELabs may join the ranks of the giants.

MELabs' Annual Stockholders Meeting was held at 4:30 p.m., June 2, 1959 at Rickey's Studio Inn, Palo Alto, California. The following directors were elected: Lloyd A. Addelman, Wesley P. Ayres, Hugh L. Gottfried, Jack L. Melchor, Perry H. Vartanian, Jr. The stockholders voted to split the stock 10 for 1 and to change the official corporate name from "Microwave Engineering Laboratories" to "MELabs."

A review of the corporate financial position for the first three quarters of the fiscal year showed a $2,535,747 backlog consisting of 17 contracts or major purchase orders. Seven of these are prime contracts. Two others are subcontracts exceeding a quarter million dollars each.



A MELabs engineer shown with his broadband ruby maser experimental setup.

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LOW POWER
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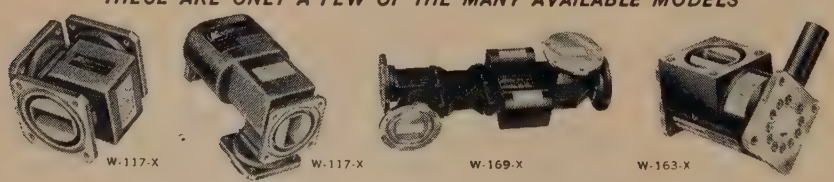
	TYPICAL SPECIFICATIONS			
MODEL	FREQ. RANGE	ISOLATION	INSERTION LOSS	V.S.W.R.
W-169-7B-5	8500-9600 MC	25 DB Min.	0.3 DB Max.	1.10 Max.
W-163-1C-1	9.2-9.4 KMC	20 DB Min.	0.5 DB Max.	1.25 Max.
W-117-2A-1	9200-9400 MC	18 DB Min.	0.3 DB Max.	1.30 Max.
W-269-2A-1	5.975-6.425 KMC	20 DB Min.	0.7 DB Max.	1.25 Max.
W-569-1B-2	16.0-17.0 KMC	I-III 30 DB Min. II-I 20 DB Min.	0.5 DB Max.	1.20 Max.

THESE ARE ONLY A FEW OF THE MANY AVAILABLE MODELS



W-117-X W-117-X W-169-X W-163-X

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CIRCLE READER SERVICE CARD 22

NEW PRODUCTS

FXR ANNOUNCES NEW POWER METER

FXR, Inc., of Woodside, N.Y. a nounces a new instrument which me ures the power of microwave signals an element only fifteen thousandths of inch in diameter.

The instrument is described by Jo E. Ebert, chief microwave engineer a a vice president of the company, as temperature compensated power meter sensitive it is accurate even in the microwatt power range. The instrumen accuracy and operating range indica the extremely high degree of sensitiv demanded of all equipment used w today's missiles and other astronauti systems, it is said.

The power measurements are made calibrating the heat generated in the ha breadth measuring element by the micr wave signal. A second and identical e ment measures only ambient or surrour ing temperatures, which produce a spu ous heating of the measuring eleme The circuitry of the instrument compe sates for the effects of ambient tempe ture variations. The result is a field strument with the incredible sensitiv and accuracy found, usually, only in la oratory standards.

Temperature compensation in this a plication of microwave instrumentation a radical and significant improvem over the methods in general use, Eb states.

CIRCLE READER SERVICE CARD 1

• • •

BROAD PULSED RF SPECTRUM ANALYSIS NOW AVAILABLE IN ONE UNIT

Featuring 200 cps resolution, the cently announced Panoramic Spectr Analyzer Model SPA-2 makes it possi to examine extremely narrow band (1 or less) microwave spectra in fine deta Half and quarter power points, lobe sy metry, and other vital characteristics broad pulsed RF are reliably delineat scan after scan.

High stability and low residual h are other outstanding features of t new Panoramic single-package unit. T versatile SPA-2 is designed for analysis standard FM and AM systems, no

"Gain"

Perhaps you first thought of bathroom scales or the waistline. Perhaps of profit.

We are more concerned with gain measurement involved in antenna development.

You might well gain — profit, that is — by using our experience in prototype and production work with plastics.

• **Fact File describes our facilities**

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In the field of microwave engineering, Wheeler Laboratories occupies a unique position as an independent laboratory devoted exclusively to consulting and development.

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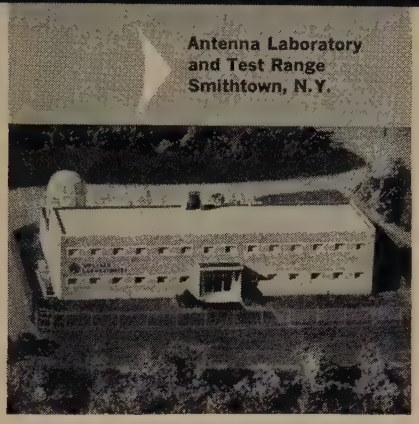

As a service to many prominent laboratories and manufacturers, WL has designed antennas, waveguide components, and microwave assemblies. Your inquiries are invited.

Harold A. Wheeler
and
Engineering
Staff

CIRCLE READER SERVICE CARD 26

New Products (Continued)

spectra, oscillator instabilities, and many other dynamic phenomena.

Other Panoramic features offered by Model SPA-2 include: 1 mc wide sweep-width, continuously adjustable down to 0 — variable resolution (10 kc to less than 200 cps) — variable sweep rate (1 cps to 60 cps) — linear, 40 db log, and power amplitude scales — choice of two extremely stable tuning heads: RF-5 (50-250 mc) —100 dbm sensitivity; and RF-6 (220-4000 mc) with —100 dbm sensitivity on fundamentals.

CIRCLE READER SERVICE CARD 104

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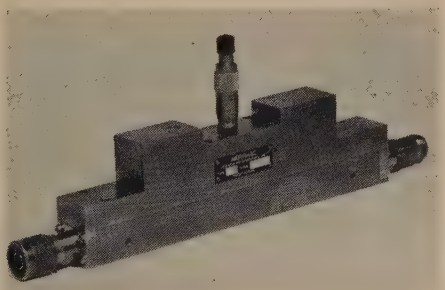
"ZERO" INSERTION LOSS EXTREMELY WIDEBAND COAXIAL VARIABLE ATTENUATOR



The Microwave Division of Merrimac Research and Development, Inc. of Flushing, New York, has recently developed a series of "zero" insertion loss coaxial variable attenuators. The new attenuators provide continuous and smooth reflectionless control of attenuation over extremely wide frequency bands and are capable of attaining very high maximum attenuation values.

Specifications are as follows:

frequency range:	300 mc to 5000 mc
input VSWR:	less than 1.35
output VSWR:	less than 1.35
insertion loss:	less than 0.5 db
power dissipation:	10 watts max.
fittings:	type N
length:	10 inches including connectors

The new attenuators are capable of providing attenuation values in excess of 40 db over most of the band. Designed to permit calibration, the attenuators are resettable to within 0.1 db.

CIRCLE READER SERVICE CARD 105

• • •

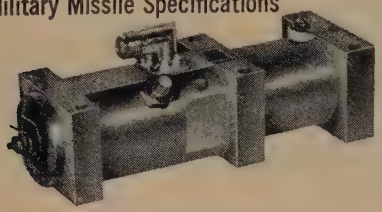
HIGH TEMPERATURE PULSE TRANSFORMER

Stavid Engineering, Incorporated, Plainfield, New Jersey, announces the development of a High Temperature Pulse Transformer. Designed for an ambient temperature of 85°C, the new device will withstand 135°C operating temperature and 150°C non-operating. The transformer uses high-temp. insulated wire with teflon pads between windings, and

NEW!

S-Band *System-Tested* MICROWAVE COMPONENTS

Engineered to meet complete Military Missile Specifications



S-BAND PULSE OSCILLATOR

- Minimum 1.5 KW Peak Power Output at .001 duty cycle; 1 μsec pulsewidth 1000 pps
- 300 MC Tuning Range
- Operating Conditions: 3000V at 2.5 amp peak (nominal)

S-BAND CW OSCILLATOR

- Minimum 50 mw Power Output 100 mw at lower frequencies
- Exceeds 400 MC Tuning Range
- Operating Conditions: 125V-250V at approx. 20-35 ma depending on plate voltage.

Both oscillators are lightweight (only 20 oz.) with integrated mountings. Both use 6442 triode tube. Heater 6.3V AC or DC.

S-BAND FILTER-MIXER-DIPLEXER

- Mixer and Diplexer (non-tunable) operate as an integral part of the Filter
- Unit features minimum Insertion Loss, Rejection Bandwidth and Input VSWR.
- Tuning Range: exceeds 300 MC

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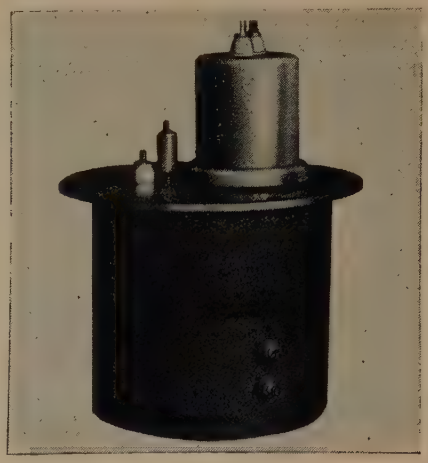
the entire unit is vacuum-impregnated with silicone oil. All gaskets and seals are silicone rubber or teflon.



Specifications are as follows:

Voltage Output:	27 KV
Step-up Ratio:	1:5
Primary impedance:	20 ohms
Pulse duration:	0.9 μsec.
PRF:	1300 pps.
Filament supply:	3.5 amps.
Weight:	14 lb.
Case diameter:	5½ inches
Case height:	5⅝ inches
Terminal:	3¾ inches

CIRCLE READER SERVICE CARD 106

• • •

MOBILE TRACKING ANTENNA

A unique new tracking antenna which combines maximum economy with utility is now available. The Transportable Tracker has three important features: a folding 28-foot reflector, hand-operated elaz drive system, and self-contained trailer mount and tower.

The reflecting surface is a special reflective cloth which folds compactly, and the ribs of the reflector are hinged around the center section. When the reflector is collapsed into a compact bundle and the supporting pillar lowered onto the flat trailer bed, the antenna is easily towed to another location by a light truck or station wagon.

While the elevation and azimuth drives are intended to be operated by hand, motors and selsyns can easily be added if desired.

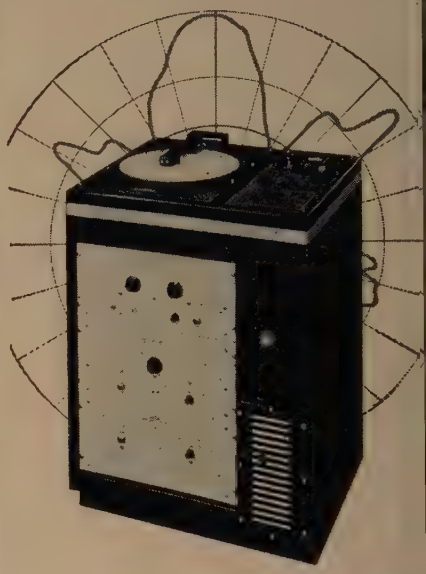
This new antenna is designed and manufactured by D. S. Kennedy & Co., Cohasset, Mass.

CIRCLE READER SERVICE CARD 107

(*More* NEW PRODUCTS *on page 54*)

CIRCLE READER SERVICE CARD 27

POLAR PATTERN RECORDING SYSTEM



Measures Antenna Radiation Patterns

AIL's Type 116R Polar Pattern recording system precisely records directional characteristics of electromagnetic, acoustic, thermal, atomic, and light radiation patterns. It traces, in polar coordinates, a continuous inked plot of input voltage vs angular position.

Precision performance is assured by low internal noise, high sensitivity, accurate turntable drive and exact pen response.

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MS FLANGES
IN STOCK

THE FINEST QUALITY BRASS AND ALUMINUM STANDARD WAVE-GUIDE FLANGES in stock at Formcraft—all precision made to military standards. For rigid and flexible waveguides.

Special flanges to your specifications. Send sketch or blueprint for a quotation.

SEND FOR NEW BROCHURE #14B
DESCRIBING AND ILLUSTRATING ALL THE STOCK FLANGES

Formcraft TOOL CO.
465 S. Archer Ave. • Chicago 16, Ill. • CAlumet 5-3173

CIRCLE READER SERVICE CARD 20

SECO

TUNING RANGE
4-8 Kmc.

POWER OUTPUT
flat within
10 db,
4-8 Kmc.
at 10 mw.

SECO TYPE OD 4-8*
BACKWARD WAVE OSCILLATOR TUBE

*Gridded

1 of 9
FREQUENCY RANGES
NOW AVAILABLE IN PRODUCTION QUANTITIES

SECO TYPE OD 4-8 is a hard-glass and metal, premium-quality tube, manufactured with extreme accuracy to meet the most exacting requirements. Used by many leading laboratories. *Send for data sheet S.*

SECO
STEWART
ENGINEERING COMPANY
SOQUEL CALIFORNIA

CIRCLE READER SERVICE CARD 21

PGMTT . . . . . national symposium 1959

Our good friend Tore Anderson was busy with his camera at the recent Symposium, and scattered around this page are some of the results. Banquet pictures courtesy Harvard University. See anyone you know?

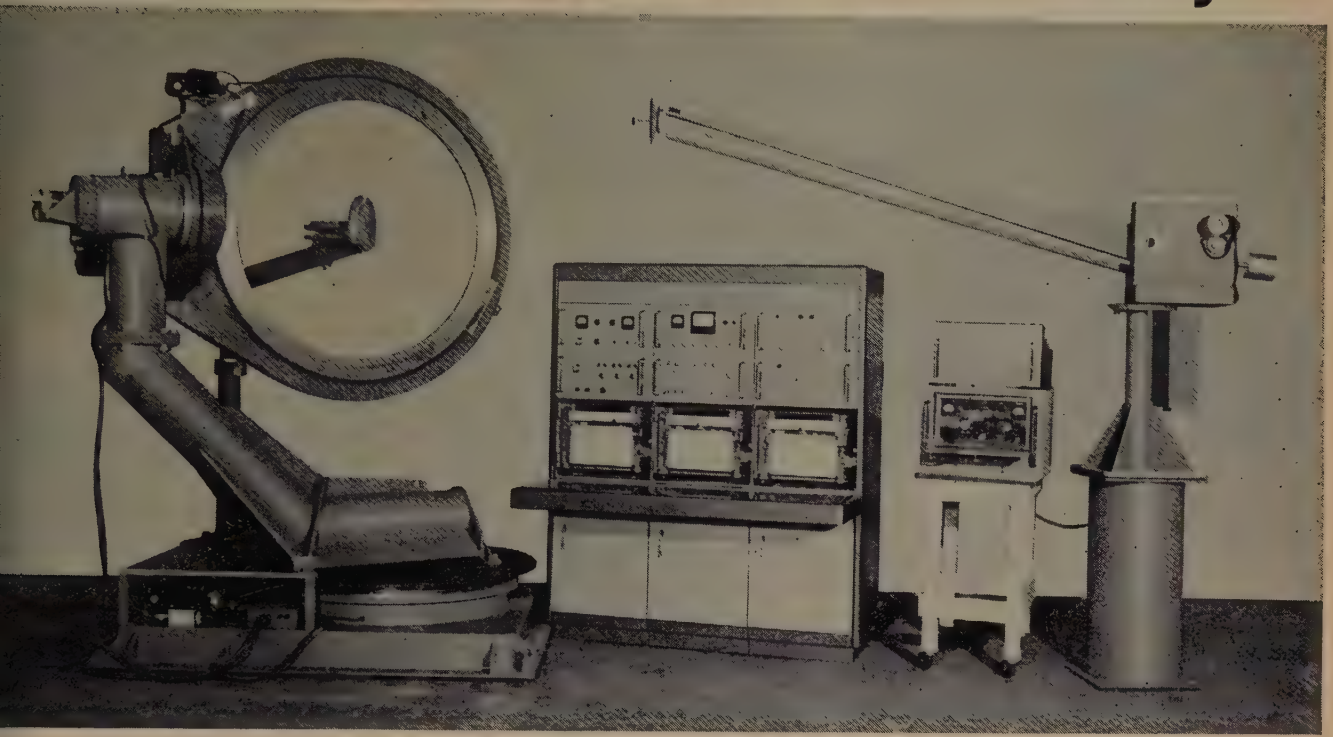
The attendance at this years' Symposium set a record for the event. Over 650 microwave people registered, and all agree it was a highly successful affair. After the only evening session, over 100 registrants visited the Cambridge Electron Accelerator for a highly interesting tour.

The banquet attendance exceeded 165. All were spellbound by Prof. E. M. Purcell of Harvard University, the main banquet speaker, who gave his own ideas on how man should approach the space age.

Credits for such a successful meeting should be extended to the 1959 PGMTT Symposium committee, headed by Bill Pritchard; Harvard University and its staff for providing such delightful facilities for the event; the local manufacturers representatives who furnished the coffee and doughnuts for the breaks. "The microwave journal" provided a slide projector operator, for whom at times the excellent acoustics of Paine Hall were no help at all. The Local Arrangements Committee takes credit, after the fact, for the variety of weather, ranging through sunny, and crisp, rainy and miserable, hot and muggy.

Next year, San Diego!

what is your *BORESIGHT* accuracy?



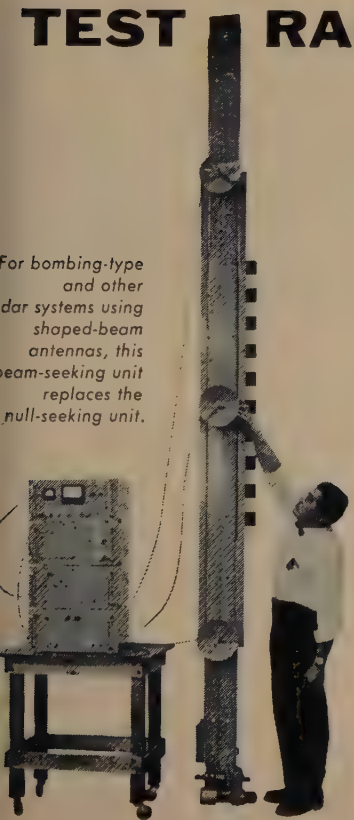
TEST RADOMES TO MIL-R-7705A (ASG)

• including type Π radomes • automatic recording of boresight error with overall system accuracy of ±0.1 milliradian ($\pm\frac{1°}{175}$)

The effectiveness of the best airborne radar tracking system can be completely destroyed by deflection of the radar beam as it passes through the plastic radome. Now, for the first time, radome manufacturers and aircraft firms installing radomes can be completely assured of radome quality. CTI Automatic Radome Boresight-Error Measuring System replaces tedious spot checking with a continuous recording of beam shift across radome.

In the system shown above, as the motor-driven holding fixture rotates the radome under test, the servo-controlled null-seeking antenna on the boom traces the position of the deflected radar beam. At the control console, which provides single-point control of the entire system, three recorders plot the total magnitude and horizontal and vertical components of the beam deflection angle directly in milliradians. Rate of change of beam deflection, essential to missile applications, is readily obtainable from the recordings.

Additional features include automatic plotting of antenna patterns with or without the radome and radome transmission efficiency measurements. The system is available in frequencies from S through K_a bands and infrared.



For bombing-type and other dar systems using shaped-beam antennas, this eam-seeking unit replaces the null-seeking unit.

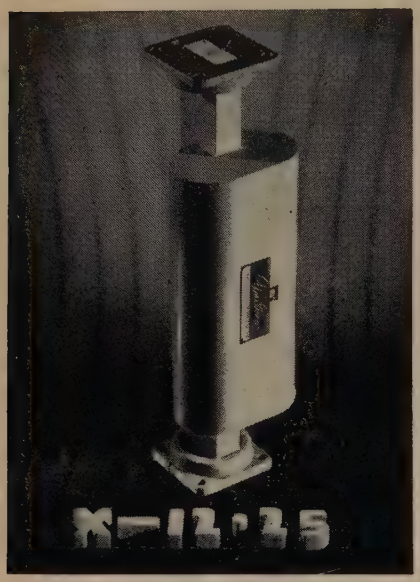
Engineers: Career opportunities are currently available at CTI

CTI®

CALIFORNIA TECHNICAL INDUSTRIES
DIVISION OF TEXTRON INC.
BELMONT 3, CALIFORNIA
Foremost in Automatic Testing

new from Cascade...

BROADBAND FERRITE LOAD ISOLATORS



designed for test equipment applications

SIX MODELS

kmc	isolation db	Model	Price
8.2—12.4	30	X—12•29	$220.
8.2—12.4	40	X—12•25	270.
7.0—10.0	30	XL—12•10	245.
7.0—10.0	40	XL—12•7	295.
5.85—8.2	30	J—12•70	220.
5.85—8.2	40	J—12•65	270.

GENERAL SPECIFICATIONS

Power capability, watts, forward	10
Power capability, watts, absorbed	2
Insertion loss, db max	1.0
Input vswr into flat load, max	1.15

Send now for additional details on these Uniline Isolators, or get in touch with us on any problems you have concerning microwave ferrite devices.

CASCADE RESEARCH

5245 san fernando rd. west
los angeles 39, calif.
chapman 5-8625

division of **MONOGRAM PRECISION INDUSTRIES, INC.**

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You'll enjoy living in this suburban town — only 47 minutes from New York City.

Send your resume today — it will be held in the strictest confidence.

Mr. Donald M. Gerber

TOP FLIGHT POSITION AVAILABLE FOR

SR. PROJECT ENGINEER

A rewarding present and attractive future awaits the man we are looking for — especially if you've had nificant or closely related experience in the followin

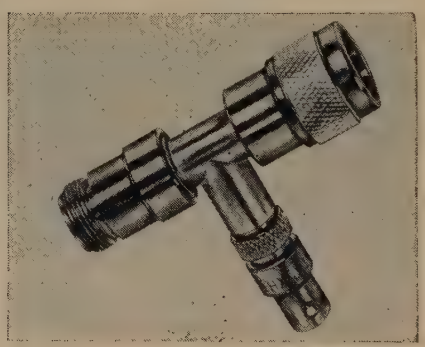
HIGH POWER PULSE MODULATOR DESIGN
- For radar and microwave systems as well as tube and components testing.

HIGH VOLTAGE POWER SUPPLY DESIGN
- To meet modern demands for TWT Klystron, Carcinotron and backward wave oscillator applications.

MANSON LABORATORIES, Inc.
375 Fairfield Ave., Stamford, Conn
DAvis 5-1391

NEW PRODUCTS

COAXIAL SIGNAL SAMPLERS FROM T. E. M., INC.



T. E. M., Inc. of Livingston, New Jersey, has recently made available three series of Coaxial Signal Samplers. Designed to monitor power in coaxial systems, the samplers are in the form of a coaxial tee. The main line consists of a section of 50-ohm line, fitted at either end with Type N, BNC, TNC, C, or HN connectors. The auxiliary, or monitoring line, is equipped with either an electrostatic- or loop-type probe whose insertion depth can be adjusted to provide a coupling variation range of at least 60 db. The auxiliary line is fitted with a female BNC connector.

The HX-series makes use of a loop probe and a self-contained crystal holder in the auxiliary line. The HY series is identical to the HX except that the crystal holder is omitted. The HZ series employs an electrostatic probe and thus provides no D-C return.

CIRCLE READER SERVICE CARD 108

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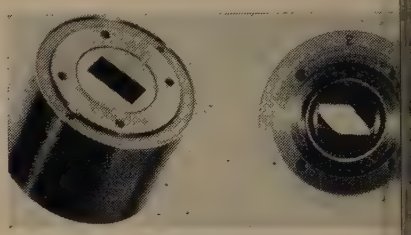
KU-BAND ROTARY STEP TWIST JOINT

Stavid Engineering, Inc., Plainfield, N.J. offers this novel step twist rotary joint which provides a reliable means transmitting Ku-Band microwave ene at high-power levels through oscillat trunnions, such as those used in scann radar antennas. The unit incorpor integral stops to eliminate the possibi of damage to the gear train from ex sive angular rotation.



Electrical characteristics are as follo Frequency range: 15. to 17.2 kilome cycles; VSWR: less than 1.09:1 o entire Ku Band; Peak Power Handl Capacity (Pressurized): 100 kilowa Insertion Loss: 0.1 db maximum.

Mechanical specifications: Rotary Mc ment: 180°; Weight: (steel)—14 oun (aluminum)—7 ounces; Dimensic 1⅞ x 1⅞ inches; Pressurization: 25 at sea level; Torque: 22 in/ounces; A bient Temperature: 55° to 100 Tested Life: 200,000 cycles at 180 total movement; Humidity: to 100 relative, at 50°C. The basic design also be scaled to S, C and X-Band.

CIRCLE READER SERVICE CARD

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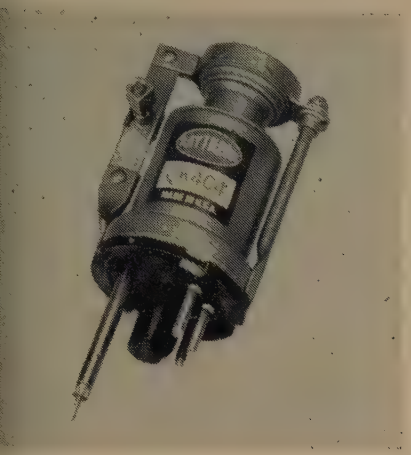
RAYTHEON GUARANTEES COMMUNICATION KLYSTRONS F 7,500 HOURS' SERVICE LIFE

Advanced manufacturing techniq and rigorous quality control procedu now make it possible for Raytheon Ma facturing Company, Waltham, Mass., increase the life warranty of its compl line of low-power communication k strons to 7,500 hours. Reports from field indicate that tubes have perform effectively for as much as 20,000 hou

Designed primarily for microwave c munications equipment, these mech ically tuned, reflex-type klystrons co government, studio link and common c

er frequency bands. Two tubes are used
each relay unit: one in the transmitter;
ne in the receiver as local oscillator.
his makes it unnecessary to maintain a
onstant relationship between outgoing
nd incoming frequencies. Tubes can be
riven in parallel from a single receiver-
pe power supply.



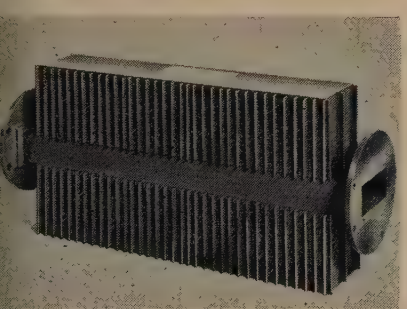
Tubes have zero temperature coeffi-
ent. No blowers are required. Tubes
re provided with a coaxial-type output
hich couples into an appropriate size
aveguide by means of a suitable trans-
ucer.

CIRCLE READER SERVICE CARD 110

• • •

HIGH-POWERED HARMONIC FILTER ANNOUNCED BY G.E.

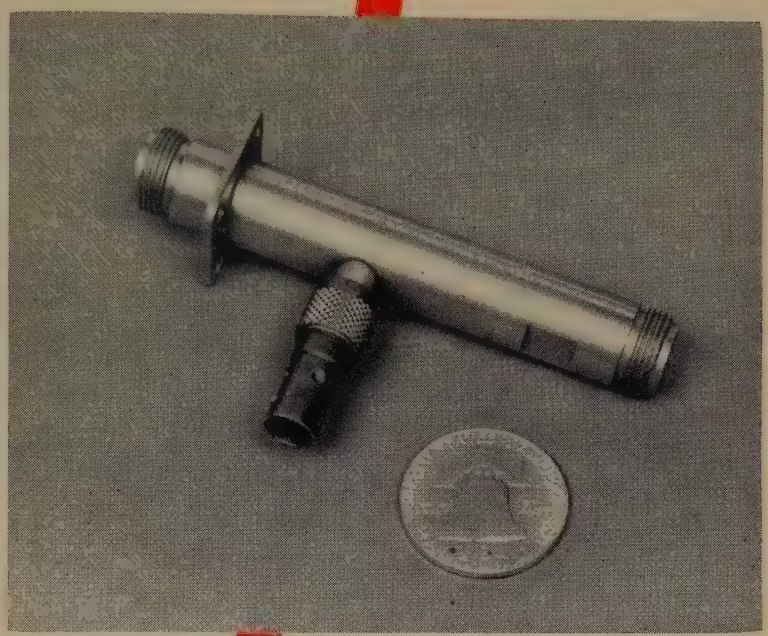
The development of a high-power har-
onic suppression filter for S-band radars
as been announced by the General Elec-
ic Power Tube Department.



Ruggedly constructed to meet the per-
rmance standards of military applica-
ons, the filter is designed to suppress
e typical harmonic and spurious signals
nerated by high-power magnetrons,
ystrons, and traveling-wave tubes.
The filter is capable of handling power
vels up to 5 megawatts. It has an in-
rtion loss below 0.2 db throughout the
ss band 2,700-3,100 MC, and an inser-
on loss in excess of 25 db for all signals
the range 5,400-10,000 MC. The
SWR across the pass and stop bands is
ss than 1.8.

IRCLE READER SERVICE CARD 111

(More NEW PRODUCTS on page 57)



SAGE

The SAGE Model 215 single ended mixer operates over the frequency range from 10 MC to 15,000 MC. Typical* sensitivity over this range is —80 to —100 dbm. The Model 215 may be obtained with or without mounting flange. This unit uses the 1N358A crystal.

Price in small quantities (less crystal) is $150.00 each.

Delivery is from stock.

*This means some units just squeeze by.

CIRCLE READER SERVICE CARD 32

SAGE laboratories, inc.
159 LINDEN STREET WELLESLEY 81, MASS.

ENGINEERS AND PHYSICISTS R & D ON NEW MICROWAVE DEVICES

Microwave Associates, Inc. invites qualified engineers and physicists to discuss an unusual opportunity to perform R & D work on new microwave devices utilizing ionized media, ferrites and semiconductors. A broad background can be achieved in the interaction of electromagnetic energy with non-linear lossy dielectrics. Here is an opportunity to grasp the advantages of each medium, combine their properties, and develop new devices.

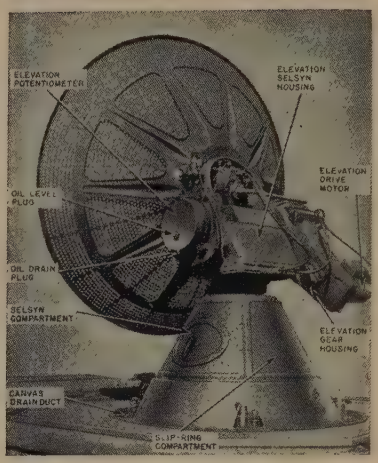
Excellent high power and low power facilities are available for complete study and test of these devices over a wide frequency spectrum — 400 to 70,000 MC.

You are invited to a confidential exploratory discussion of your future with us. Contact Dr. Lawrence Gould for suitable arrangements.

MICROWAVE ASSOCIATES, INC.

BURLINGTON
MASSACHUSETTS
Telephone
BRowning 2-3000

RADIO RESEARCH INSTRUMENT CO.
550 FIFTH AVE. NEW YORK JUDSON 6-4691



ANTENNA PEDESTAL SCR 584—MP 61B

Full azimuth and elevation sweeps. 360 degrees in azimuth. 210 degrees in elevation. Accurate to 1 mil. over system. Complete for full tracking response. Includes pedestal drives, selsyns, potentiometers, drive motors, control amplidynes. Excellent used condition. This is the first time these pedestals have been available for purchase. Limited quantity in stock for immediate shipment. Ideal for antenna pattern ranges, radar systems, radio astronomy, any project requiring accurate response in elevation and azimuth. Complete description in McGraw-Hill Radiation Laboratory Series, Volume 1, page 284 and page 209, and Volume 26, page 233.

AUTOMATIC TRACKING RADAR

Complete system for missile, aircraft, satelite, balloon tracking to altitude of 100 miles or better. Housed in own van 20 ft. x 12 ft. x 8 ft. As new. Write us.

2 MEGAWATT PULSER

Supplies 2 megawatt to magnetron at a pulse duration of 2 microsec. Fixed gap type. Uses 4C35 hydrogen thyratron. Brand new complete. G.E. $1150.00.

F-28/APN-19 FILTER CAVITY

Jan. spec: Tuneable 2700-2900mc, 1.5db max. loss at ctr freq over band. Details: Insertion loss variable. Single tuned filter for freq channelling in radar beacon. Invar center tuning conductor ¾ wavelength. New $37.50 each.

⅞" RIGID COAX. RG44/U 50 ohm, standard fittings, 10cm stub supported. 12 ft. lengths. Silver plated. New $34.50 each. 12 ft. length. Right angle bends $6 ea.

KLYSTRON MOUNTS

3CM. Precision Tube Mount. Waveline model 688. X band shielded klystron mount PRD signal generator type. Complete with variable glass vane attenuator. Brand new. $205. list. Price $45.00.
S. band. Type N output. Tunable over entire band. For Shepard type tube i.e. 726 w/socket & tube clamp. Mfg. GE. New, $15.00.

TOPWALL HYBRID JUNCTION. 8500-9600mc 1x.5 wg size. Broad banded better than 16%. Aluminum casting. $15.00 new. Crossover output. 1x.5 wg size. $5.00 new.
BROAD BAND BAL MIXER using short slot-hybrid. Pound type broad band dual balanced crystal holder. 1x.5 wg size. $25.00 new.

VD-2 PPI REPEATER

Takes any radar PPI info and displays on 7" tube compact floor standing console. Complete rotating yoke assy. 115V 6DCY AC Pwr. 4, 20, 80, 200 mile ranges. Brand new w/Inst. Bk. $385.

AN/APS-10 3CM. X BAND RADAR

Complete RF head including transmitter, receiver, modulator. Uses 2J42 magnetron. Fully described in MIT Rad. Lab. Series Vol. I, pps. 616-625 and Vol. II, pps. 171-185. $375.00. Complete X band radar system also avail. incl. 360 deg. antenna, PPI, sync, pwr supply. Similar to $17,000 weather radar now in use by airlines. $750 complete.

CIRCLE READER SERVICE CARD 33

RESEARCH ENGINEERS

Challenging assignments are presently available for Electronic Engineers at all levels of experience to work on interesting military and industrial projects in the following areas:

Communications Systems - Radio-Radar Systems

Electronic Instrumentation

Electronic Components - Electronic Interference

Microwave and Propagation Studies

Members of our Electrical Engineering Research staff enjoy an exceptional opportunity for personal and professional growth through project assignments carried out in an atmosphere conducive to good research. In addition they receive competitive salaries and benefits which include a generous relocation allowance and up to four weeks paid vacation.

If you are seeking increased responsibility and professional recognition, send a resume of your qualifications to:

A. J. PANERAL

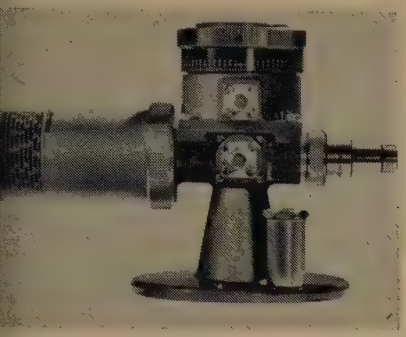
ARMOUR RESEARCH FOUNDATION

of Illinois Institute of Technology

10 West 35th Street · Chicago 16, Illinois

NEW PRODUCTS

NEW PRD COAXIAL STANDING WAVE DETECTOR



A coaxial Standing Wave Detector that ermits direct reading of angle of reflec- on coefficient was recently announced Polytechnic Research & Development ., Inc., Brooklyn, N.Y.

Designed to obsolete huge 8-foot units, e PRD Type 219 Standing Wave De- ctor measures impedance and VSWR om 100 to 1000 mc/s.

In addition, PRD's rugged, low-cost ype 219 provides direct reading of SWR; direct reading of reactive compo- ent sign; adaptability to most coaxial ne, including the LT and the new TNC ries. A matched load for self-calibra- on is also supplied.

Important specifications of the PRD ype 219 include: residual VSWR of ss than 1.03; minimum input signal of proximately 1 V at 100 mc/s, 0.1 V 1000 mc/s for measuring a matched ad; and characteristic impedance of 50 ms.

IRCLE READER SERVICE CARD 112

• • •

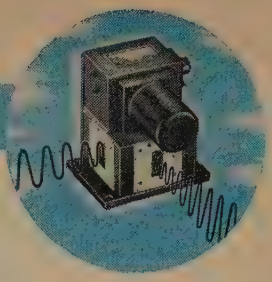
RAYTHEON C-BAND ISOLATORS

Raytheon now offers three medium ower C-band isolators covering the com- ercial band.

They are the Raytheon Model Nos. CM13 (5925-6475 Mc/s), ICM12 (6575- 25 Mc/s) and ICM21 (7125-7650 c/s).

Each of these isolators has UG-344/U nges, mates with RG-50/U waveguide, eighs 2.6 pounds and is 5.0 inches long; us complete interchangeability is as- red.

PROGRESS REPORT



Microwave wave-guide switches

In the state-of-the-art of wave-guide switches, TAPCO Group microwave engineers in 1955 pioneered the perfection of switches capable of transferring from one band to another under full power, without shutting down the transmitter by interlocks. This development involved S and L band switches which operate over their respective wave-guide frequency ranges with no tuning required. Power tests have been conducted on these units, unpressurized, with the S band switch transmitting and switching under 4.6 megawatts peak pulse power and 4.6 kilowatts average power; and the L band switch transmitting and switching under 9.8 megawatts peak pulse power and 9.8 kilowatts average power. Each of these tests was limited only by the power source available, and not by the performance of the switch.

An extension of this switching principle has resulted in a unique wave-guide-switch-and-power-divider unit which, in its present form, is capable of switching full wave-guide power to either one of two output wave-guide lines or of dividing the power equally between these lines.

Other units can be built by the TAPCO Group to give any selected power split between the two output lines up to the crosstalk value of the basic switch design. Additional possibilities would be a unit capable of several stepped values of division, or a unit driven at a constant speed and programmed externally to desired power split values.

The first single-pole, two-throw wave-guide switch for double-ridged wave-guide operating in the frequency band from 4750 to 10,500 mc/s was also developed by TAPCO Group microwave engineers. This unit could also handle the full wave-guide power with insertion VSWR of less than 1.15/1 and crosstalk greater than 70 db.

A unique, single-pole, four-throw wave-guide switch recently developed by the TAPCO Group is probably the first high crosstalk switch of this type available for microwave systems. It is designed to carry full X-band peak and average powers over the entire wave-guide frequency band of 8.2 to 12.4 kmc/s, with more than 90 db crosstalk rejection and a VSWR of less than 1.06/1.

Other microwave components currently under development at TAPCO Group include microwave electronic counter-measure antennas; power dividers; non-contacting L-band lobe switches for long life, service-free IFF systems; and other transmission line subsystems.

Further information on the capabilities and facilities of the TAPCO Group in the development and production of microwave systems and components will gladly be sent you on request.

TRW

TAPCO GROUP

Thompson Ramo Wooldridge Inc.

Dept. MJ-759 • Cleveland 17, Ohio

CIRCLE READER SERVICE CARD 34

Microwave Engineers

Join long-range, advanced development programs to create new microwave devices

In the Microwave Development Laboratories of the Radio Corporation of America, at Princeton, N. J., and Lancaster, Pa., engineers are engaged in long-range programs on a variety of microwave devices (super power klystrons, grid controlled amplifiers and switch tubes, conventional power level microwave devices, traveling wave tubes, parametric amplifiers and microwave solid state devices, high-speed switches and high perveance electron guns).

In essence, the challenge in the advanced development laboratories is to achieve further advances in these areas of significant scientific importance.

Opportunities are available offering unlimited progress to engineers with PhD and MS degrees (Physics and EE) who can contribute new concepts in these fields or can lead engineering groups investigating new microwave devices and utilizing new operating principles.

INVESTIGATE THESE OPPORTUNITIES — SEND RESUME IN COMPLETE CONFIDENCE:

FOR PRINCETON, N. J.

Mr. J. F. McPartland
Professional Personnel
Dept. J-3G
Radio Corporation of America
Electron Tube Division
Harrison, N. J.

FOR LANCASTER, PA.

Mr. E. F. Stevens, Mgr., Employment
Dept. G-15G
Radio Corporation of America
Electron Tube Division
New Holland Pike, Lancaster, Pa.

EQUAL JOB OPPORTUNITY For All Qualified Applicants

RCA

RADIO CORPORATION of AMERICA
ELECTRON TUBE DIVISION

Rated at 5 watts average power, th 40 db minimum isolation is desig especially for commercial systems to duce intermodulation distortion to a n ligible level with a maximum inserti loss of 1 db and a VSWR of no m than 1.15 over the band.

These devices are readily availa from Raytheon Company, Special Mic wave Devices Subdivision.

CIRCLE READER SERVICE CARD 1

LITERATURE . . .

G.E. BULLETIN OFFERED

General Electric's new "Dual-Bea Microwave equipment is described Bulletin ECM 91, available from G Communication Products Departme Lynchburg, Va.

The bulletin explains how the "Du Beam" concept offers additional prot tion against unpredictable communi tion failures by exceeding techniques us ally found in automatic standby equi ment. Cross-polarization of antenn makes continuous parallel operation two sets of RF equipment possible, th increasing reliability of the total syste

CIRCLE READER SERVICE CARD 1

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NEW LITERATURE FROM PRD

The latest issue of "PRD Reports," V 6, No. 1, written by Dr. Samuel Hopf and Leo Nadler of the staff of the Pol technic Research and Development C Inc. is entitled "Waveguide Rotary Stan ing Wave Indicators." This six pa technical paper describes the theory, o eration and design of RSWI's.

These new measuring devices, call Rotary Standing Wave Indicators, prese a bold solution for VSWR and impedan measurements in waveguide syste RSWI's replace slotted sections whe length, size, and weight are importa factors.

CIRCLE READER SERVICE CARD 1

• • •

PARABOLIC ANTENNA SYSTEM COMPUTER AVAILABLE FROM ANDREW CORPORATION

The Andrew Corporation, manufactur of parabolic antennas and microwa transmission line systems, has recent revised its Parabolic Antenna Syst Computer and Transmission Line Wav guide Selector.

The system computer enables commu cation and microwave system engineers rapidly and accurately calculate parabo antenna radiation characteristics, passi repeater performance, free space a "Scatter Propagation" attenuation, a thermal and equivalent noise input receivers.

The reverse side of the computer is Transmission Line and Waveguide S lector on which attenuation and pow ratings, as well as other electrical a mechanical characteristics, are shown.

CIRCLE READER SERVICE CARD 1

PHILCO Silicon Microwave Mixer Diodes

Offer Unequalled Performance and Sensitivity...at 16,000 mc and 24,000 mc

- Lowest Over-all Receiver Noise Figure
- Operating Temperature More Than Doubled
- Hermetic Seal for Maximum Reliability
- Burn-Out Rating Doubled

Performance Data

	1N26	1N26A	*1N26B	
Conversion Loss	8.5	7.5	7.5	(db)
Noise Ratio	2.5	2.0	1.5	(times)
RF Impedance	—	1.6	1.5	(VSWR)
Over-all Receiver Noise	13.1	11.3	10.0	(db)

	1N78	1N78A	1N78B	*1N78C	
Conversion Loss	7.5	7.0	6.5	6.0	(db)
Noise Ratio	2.5	1.5	1.3	1.3	(times)
RF Impedance	—	1.6	1.6	1.5	(VSWR)
Over-all Receiver Noise	11.8	9.8	8.8	8.3	(db)
Burn-Out	0.3	0.3	0.3	0.6	(ergs)

The result of extensive research and development by Philco, the two newest diodes in these families provide better performance, greater reliability and increased sensitivity in both the 16,000 mc and 24,000 mc regions. Operating temperatures of all diodes in both families are more than doubled . . . to 150° C. All Philco Diodes are packaged in hermetically sealed cases for extreme reliability under adverse environmental conditions. The newest diodes (1N26B, 1N78C) offer the lowest over-all noise figure in these series . . . immediately available only from Philco. If you are designing for maximum performance and sensitivity, to meet rigid specifications . . . write for detailed data and application information on these new Philco Silicon Microwave Diodes.

We offer complete production capability for classified millimeter crystals and invite your inquiry.

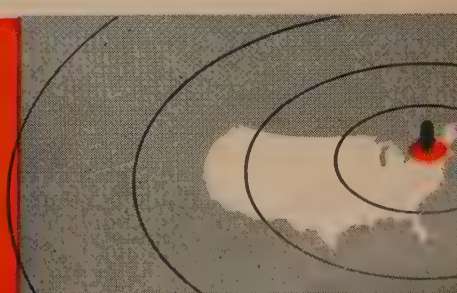
*Immediately available only from PHILCO

Write Special Components Dept. MJ-759—Lansdale Tube Company Division, Philco Corporation, Lansdale, Pa.

PHILCO

LANSDALE TUBE COMPANY DIVISION

LANSDALE, PENNSYLVANIA



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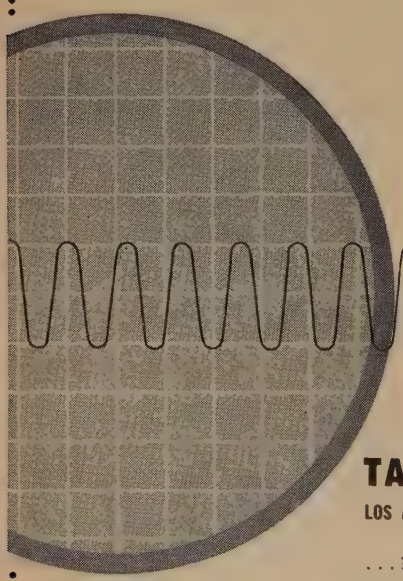

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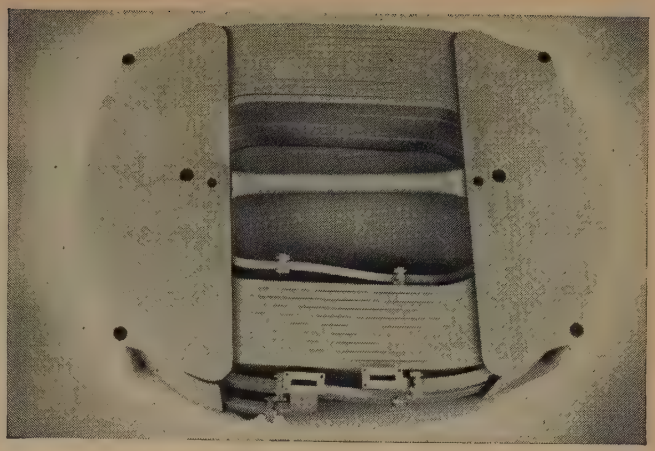

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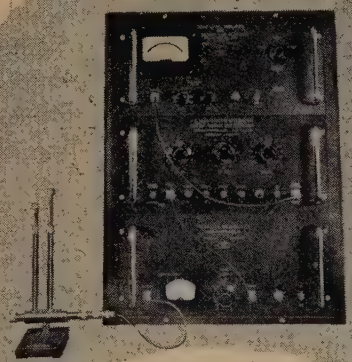

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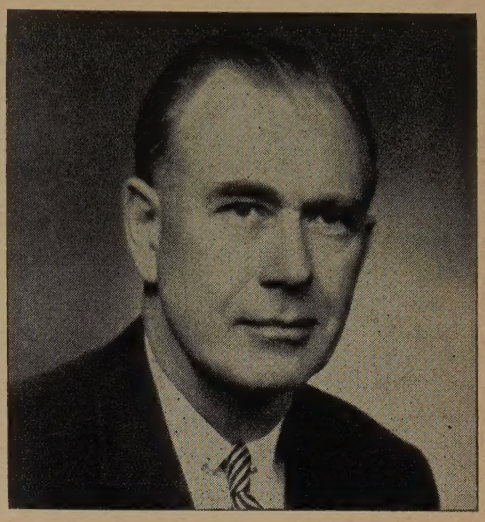

OBERT E. LEWIS, President of Sylvania Electric Products Inc., a holly-owned subsidiary of General Telephone & Electronics Corpo-ation, is widely recognized as an important contributor to the present-ay science and philosophy of American industrial management.

Prior to his election as President in December, 1958, Mr. Lewis vas a Senior Vice President of the company. He is a Director of ylvania, and also served, while a Senior Vice President, as President f the company's Argus Cameras Division and had responsibility for he operations of the Semiconductor Division. His headquarters are t Sylvania's executive offices in New York City.

Previously, as President of Argus Cameras, Inc., Ann Arbor, Aich., an independent company before its combination with Sylvania n January 2, 1957, Mr. Lewis demonstrated his management leader-hip and ability by building Argus into the world's principal producer f 35-millimeter cameras, slide projectors and related accessories.

Mr. Lewis joined Argus as Vice President and general manager n the fall of 1949, and was named President at the age of 44 the ollowing year.

His interest in the science of management led him to play a major part, over the last several years, in the activities f the American Management Association, an organization he now serves as a Director.

Mr. Lewis was elected a Vice President and Director of Sylvania, respectively, in January and April, 1957, and a enior Vice President in January, 1958.

Prior to joining Argus, Mr. Lewis was a consulting management engineer with Sanderson & Porter, world-wide ngineering firm. He was in charge of an engineering mission to Germany in 1948 on reparations for the U. S. overnment.

Earlier, he had been treasurer and a director of American Steel & Wire Co., a U. S. Steel subsidiary; president f Cleveland Wire Spring Co., and controller and assistant to the president of Batten, Barton, Durstine and Os-orn, New York advertising agency.

Born in Flushing, Long Island, New York, on September 5, 1906, he studied accounting at Columbia Uni-ersity's Graduate School of Business and held a Certified Public Accountant certificate for the states of New York nd Ohio.

Mr. Lewis is a Past President of the National Association of Photographic Manufacturers, and is Chairman of he Yale National Parents Committee. In 1935 he married the former Alice Imholz of Flushing, N.Y. They have ight children and reside in Darien, Conn.

THE SIGNIFICANT ROLE OF MICROWAVES IN THE FUTURE EXPANSION OF ELECTRONICS

By
ROBERT E. LEWIS
President
SYLVANIA ELECTRIC PRODUCTS INC.
a subsidiary of
General Telephone & Electronics Corporation

An observation heard with increasing frequency at usiness gatherings these days is that the remarkable rowth now going on within the U.S. economy, and ad-itional expansion anticipated over the next 10 or 20 ears, actually may be attributed to a Second Industrial evolution — major changes in the growth, location and operation of the various industries that make up the economy. Such comments invariably evoke from me a spirited response about the particular industry that has played, and will play, a double-barreled role in this so-called revolution. I'm talking, of course, about electronics — the industry which evolved out of Dr. Lee

De Forest's invention of the vacuum tube some 53 years ago, and whose sales and revenues have climbed from some $500 million just before the war to more than $13 billion in 1958.

It is interesting to note that this period of growth closely parallels that marked by the emergence and refinement of radar, a by-product of the microwave art, into the highly advanced radar systems we know today. The parallel is noteworthy not only because radar is among the most dramatic electronic developments of recent years, one traditionally cited by those listing significant industry milestones, but also because it symbolizes the significant role that microwaves will play in the future expansion of electronics.

Research Tools and Missile Systems

Today, microwave devices are utilized as tools of fundamental research as well as in such extremely important practical applications as radar. In the laboratory, microwaves are used in spectroscopy, particle accelerators, and in radio astronomy; elsewhere, they are employed in such diverse applications as weapons and missile systems, television relays, maritime and aircraft navigation, cancer therapy, food sterilization and home cooking ranges. In defense systems, microwave applications are both numerous and vital — including radar, communications, navigation, and the launching, guiding and fusing of missiles. A few of the accomplishments foreseen as a result of current and of future microwave research certainly would include radar steering of automobiles, anti-collision devices, and "atomic clocks" with a new and superior standard of time.

The opportunities and challenges that Sylvania foresees in the anticipated microwave expansion are best indicated by pointing out that we recently consolidated our microwave operations, including special tube and component research, development, and production. The purpose of the move, of course, was to inaugurate a vigorous program of special tube and component research, development and production planned under a unified strategic approach to the Government and industrial markets. And you can be sure this program is well under way.

Military Benefits Sparked Microwave Development

Since all of the microwave advances I mentioned have occurred in a relatively brief span of time, I invariably find myself surprised when recalling that microwave research dates back to the work of Heinrich Hertz in 1888. At that early date, he noted the similarity of electromagnetic radiation with ordinary light, and illustrated that the former, like light, adhered to the optical laws of reflection, refraction and polarization. However, since radio waves were utilized initially for long-distance communication (requiring power available only in longer wave lengths), this early investigation into microwaves was not put to practical use.

It was only in the years immediately prior to World War II, when the military advantages of "seeing" through darkness and overcast became apparent, that microwave development was pushed. Important the retical studies at the Massachusetts Institute of Tec nology and other laboratories led to discovery of t principle of guided propagation. A new and ingenio method of obtaining radio frequency power was e bodied in the klystron tube, and valuable studies England led to rapid development of the magnetro The effectiveness of the resultant radar equipment utilized for search, navigation and bombardment duri World War II — need not be reiterated.

Sylvania Enters Microwaves in World War II

One of Sylvania's earliest contributions to the micr wave component field (and a result of its extensi know-how in radio and receiving tube manufacturin was the development of microwave diodes, in conjun tion with the Radiation Laboratory at Massachuse Institute of Technology, during World War II. We we also involved in the development and manufacture cathode ray tubes for radar use during this period. Wi these activities serving as the foundation, Sylvania h expanded to become one of the industry's leaders in d velopment and production of specialized electronic sy tems, equipment and devices. In 1958, the compa spent about 6% of each non-military sales dollar, more than $15,000,000, on its own over-all research a engineering — and a sizeable slice of this was devote to electronic systems. Hundreds of scientists and eng neers are engaged in this vital work. Included amor our more important current research and engineeri projects are microwave communications, such as mas and parametric amplifiers, traveling wave tubes a ceramic tubes; and weapons and missile systems.

The bulk of the company's microwave activities a carried on within the Electronics Systems Division, wit additional microwave research — largely pioneering the upper frequency end of microwaves — under way the Sylvania Research Laboratories at Bayside, Lon Island, N.Y. Typifying the company's advanced wor in new fields such as the microwave art are the nin laboratories of the Electronics Systems Division. A Waltham, Mass., the division has its Avionics Labor tory, Missile Systems Laboratory, and Applied Researc Laboratory. At Mountain View, Calif., there are th Electronic Defense Laboratory, Reconnaissance Syste Laboratory, Microwave Tube Laboratory and Microwav Physics Laboratory. Two new laboratories opened i 1958 are the Amherst Engineering Laboratory, in Wi liamsville, N.Y., and the Data Processing Laborator in Needham, Mass.

The substantial post-war growth of Sylvania withi electronics — and particularly the microwave sector o electronics — illustrates quite well, I believe, our fait in the opportunities that exist and the challenges tha will materialize in the future — a future whose horizon have been substantially broadened by our joining th General Telephone & Electronics organization. A of this serves to indicate why I am inclined toward en thusiasm when reviewing the double-barreled role tha electronics is playing and will play in the Second In dustrial Revolution.

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BL-223	5.0	0.7	1900 V	1.1	400	5400–5900	50Ω TNC Plug	10 oz.	
BL-226	5.0	0.5	1300 V	0.9	100	9100–9500	50Ω TNC Plug	8 oz.	
BL-227	5.0	0.5	1300 V	0.9	100	8700–9100	50Ω TNC Plug	8 oz.	
BL-228	5.0	0.5	1300 V	0.9	100	8300–8700	50Ω TNC Plug	8 oz.	
BL-230	5.0	0.7	2800 V	1.9	1000	5400–5900	50Ω TNC Plug	10 oz.	
BL-231	5.0	0.5	1450 V	1.0	200	5400–5900	50Ω SM Jack	8 oz.	
BL-233	5.0	0.7	2800 V	1.5	1000	9375	UG 40/U	11 oz.	
BL-242	5.0	0.7	1900 V	1.1	400	5400–5900	TNC or 50Ω N Plug	10 oz.	
BL-243	5.0	0.5	1450 V	1.0	200	5400–5900	50Ω SM Jack	8 oz.	
BL-245	5.0	0.7	2800 V	1.9	900	5400–5900	50Ω TNC Plug	10 oz.	
BL-247	5.0	0.5	1200 V	0.8	100	9100–9500	50Ω TNC Plug	8 oz.	
BL-250	5.0	0.5	1350 V	1.0	150	5400–5900	50Ω TNC Plug	8 oz.	
BL-M003	5.0	0.5	1300 V	0.9	100	9100–9500	50Ω TNC Plug	7 oz.	
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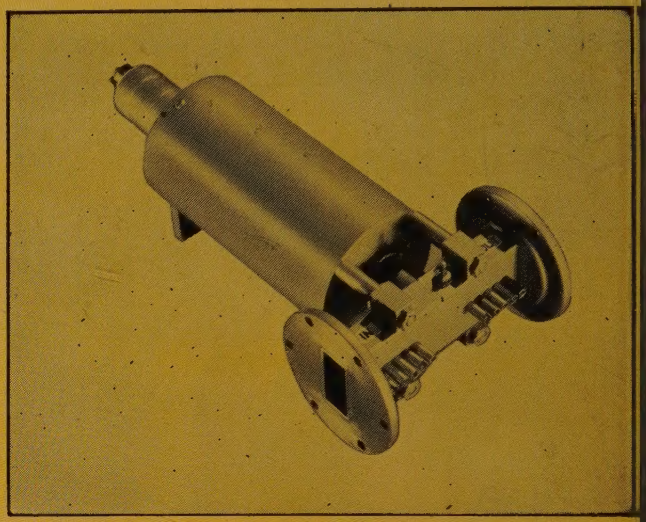
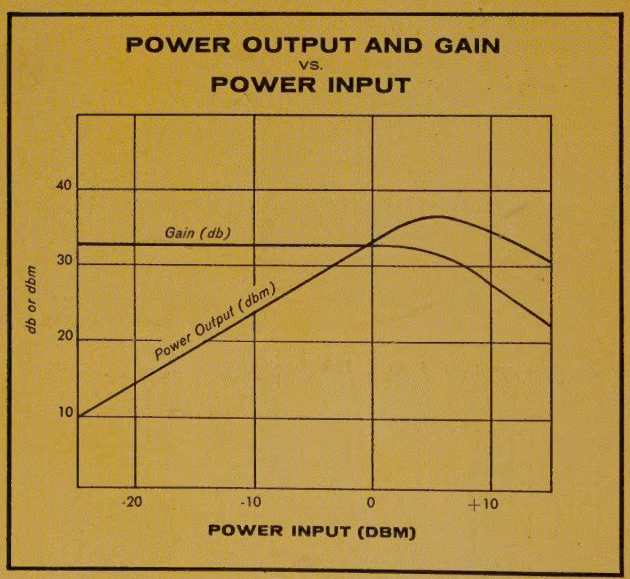
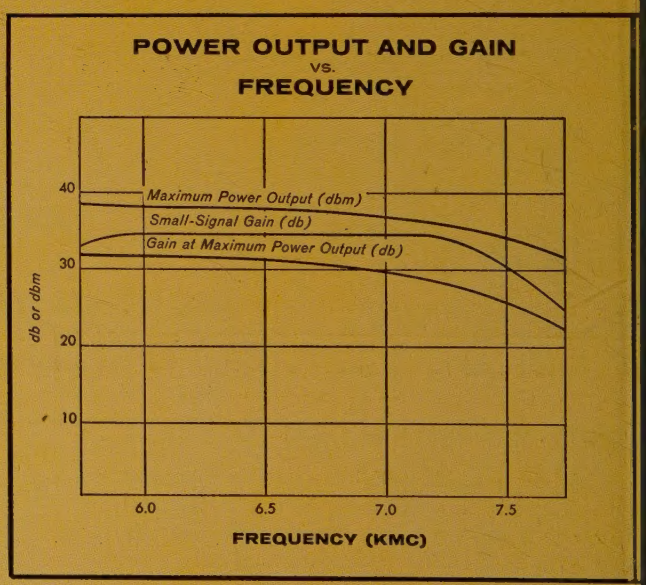


Typical Operating Characteristics

Frequency Range	5,900 to 7,400 Mc
VSWR (Input and Output)	2.1:1 max.
Small-Signal Gain	32 to 35 db
Gain (Saturation)	25 to 27 db
Power Output	5 watts





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